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DECEMBER 1978

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More Details? CHECK—OFF Page 150

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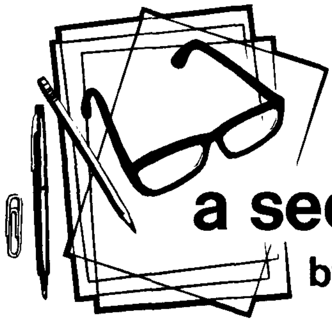

a second look

by Jim Fisk

Small-scale integration, large-scale integration, counters on a chip, 64K memories — are the possibilities limited only by the designer's imagination? I can think of no other segment of the electronics industry which has come so far, so fast, as integrated circuits. When vacuum tubes were 20 years old they were mired down by bitter patent litigation which had slowed development work to a standstill. Marconi, the Fleming patent holder, even placed a full-page advertisement in *QST* warning that amateurs who even *used* other than Marconi-licensed tubes for radio purposes were "liable to a suit for injunction . . . and they will be prosecuted." A scare tactic, perhaps, but it was effective and held back the progress of the entire radio industry.

No such patent litigation ever developed in the transistor industry because the inventors placed the basic idea in the public domain, so by the time transistors reached their 20th birthday there were more than 2000 registered types, and the industry was thriving. Circuit development, however, was stymied by engineers schooled in vacuum-tube techniques who struggled to make solid-state devices work as well as tubes. If you used any of the early solid-state ham gear you know that they were not always successful. In fact, there are some amateurs who still believe that, in a communications receiver, a solid-state front end is inherently inferior to a vacuum tube. In truth, solid-state front ends are actually several hundred times *better* than their vacuum-tube counterparts — both in strong-signal handling and sensitivity — but only when properly designed.

The rapid growth of integrated-circuit technology has been hampered neither by patent squabbles nor inept circuit design — to engineers who cut their teeth on transistors. ICs were the next logical step. So in this, their 20th year, ICs have already affected every industry, every level of society, and the end is nowhere in sight.

It was during the summer and fall of 1958 that Jack Kilby built the first integrated circuit at Texas Instruments. Other semiconductor manufacturers had been working on ways to miniaturize electronic circuits, but most of them relied on miniature discrete components. Kilby was the first to use semiconductor material for both the active and passive elements (resistors and capacitors) to build a complete circuit on a single piece of germanium. (Germanium was used because germanium manufacturing techniques were well established, and those for silicon were not.) Kilby's first circuits, a phase-shift oscillator and multivibrator, demonstrated the feasibility of this approach. On top of the germanium substrate were the contacts of the diffused transistors, junction capacitors, and resistors. A gold-plated metal frame protruded from the lower surface of the wafer and thermally-bonded gold wires were used for connections between those elements not linked by the substrate itself.

The first circuits were crude by today's standards — large and irregular; the photo masks and resists necessary for precision IC manufacturing were yet to be developed, so the patterns were hand painted on the semiconductor chip with black wax. About the same time Kilby was working on the first ICs in Texas, Fairchild Semiconductor developed the *Planar* process; this process made semiconductors more reliable and less expensive to produce, and greatly accelerated IC progress and acceptance.

In the 20 years since those early discoveries, prices have decreased and the number of circuits per unit area has increased dramatically. In 1962, for example, a typical IC flip-flop was about 2 mm square; a similar circuit in 1968 was ten times smaller, and today an entire 8-digit frequency counter with all control functions is available on a single chip. In 1962 a decade counter required several counter chips and logic gates for a total cost of twenty or more uninflated dollars; by 1968 the cost of a one-chip decade counter had dropped to about seven dollars; today you can buy a TTL decade for fifty cents! And you must remember that 1962 counter did well to count reliably at 1 MHz; the 1978 counter is guaranteed to 30 MHz.

Progress in the field of linear ICs has been just as impressive, and the many functions which are built into modern amateur transceivers are a direct result. Some of the newer transceivers have 25 or more controls on the front panel; can you imagine how many 6-foot racks of vacuum-tube circuits it would take to do the same job? And even if you had room for the racks, you wouldn't want to pay for the electric power to run them (and cool them). If progress in the next ten years is as rapid as in the past ten, the commercial equipment that will be available to amateurs in 1988 really boggles the mind.

James Fisk, W1HR
editor-in-chief

New, Remotable 2meter Mobile!

ICOM's New IC-280

ICOM introduces its new 2 meter mobile radio with the detachable microprocessor control head, the **IC-280**. Bright, easy to read LED's and a new style meter grace the brushed aluminum "new look" front panel of the detachable control head, which provides memory and frequency control for the remotely mountable main section.

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With the microprocessor head the **IC-280** can store three frequencies of your choice, which are selected by a four position front panel switch. These frequencies are retained in the **IC-280's** memory for as long as power is applied to the radio. Even when power is turned off at the front panel switch, the **IC-280** retains its programmed memories; and when power is completely removed from the radio, the ±600 KHz splits are still maintained!

Frequency coverage of the **IC-280** is in excess of the 2 meter band; and the new band plan (144.5-145.5 MHz repeaters) can easily be accommodated, since it was included in the **IC-280's** initial planning by the ICOM design team.

The main section of the **IC-280** puts you up to the minute with the latest state of the art engineering. The new **IC-280** includes the latest innovations in large signal handling FET front ends for excellent intermodulation character and good sensitivity at the same time. The IF filters are crystal monolithics in the first IF and ceramic in the second, providing narrow band capacity for today and tomorrow's crowded operating conditions. Modular PA construction with broad band tuning provides full rated power across the full 2 meter band (plus a little).

All ICOM radios significantly exceed FCC specifications limiting spurious emissions.

Specifications subject to change without notice.

IC-280 Specifications: □ Frequency Coverage: 143.90—148.11 MHz □ Operating Conditions: Temperature: −10°C to 60°C (14°F to 140°F), Duty Factor: continuous □ Frequency Stability: ±1.5 KHz □ Modulation Type: FM (F3) □ Antenna Impedance: 50 ohms unbalanced □ Power Requirement: DC 13.8V ±15% (negative ground) □ Current Drain: Transmitting: 2.5A Hi (10W), 1.2A Lo (1W), Receiving: 0.630A at max audio output, 0.450 at SQL ON with no signal □ Size: 58mm(h) x 156mm(w) x 228mm(d) □ Weight: approx. 2.2 Kg □ Power Output: 10W Hi, 1W Lo □ Modulation System: Phase □ Max. Frequency Deviation: ±5 KHz □ Spurious Output: more than 60 dB below carrier □ Microphone Impedance: 600 ohms dynamic or electret condenser type, such as the SM-2 □ Receiving System: Double superheterdyne □ Intermediate Frequency: 1st: 10.695 MHz, 2nd: 455 KHz □ Sensitivity: 1 uv at S+N/N at 30 dB or better, Noise suppression sensitivity 20 dB, 0.6 uv or less □ Selectivity: less than ±7.5 KHz at −6 dB, less than ±15 KHz at −60 dB □ Audio Output: More than 1.5W □ Audio Output Impedance: 8 ohms

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comments

transient suppression

Dear HR:

I have read the article in your June 1978, issue, "Protecting Solid State Devices From Voltage Transients," with great interest. Mr. Prudhomme has written a very informative article that apparently was well researched, but I question his statement that varistors are excellent in suppressing transients on the line side of equipment power supplies.

The surge energy capacity of the MOV is proportional to the volume of its active material, and the breakdown voltage is proportional to the thickness of its material. Page 57 of the GE transient voltage suppression manual, referenced in Mr. Prudhomme's article, addresses itself to (1) Overstress Near Ratings, and (2) Extreme Overstress. The varistor has its place in the proper environment, but its application must be carefully chosen. As pointed out in Prudhomme's article, transients cover a wide range of frequencies and amplitudes. How do you select the proper component to protect against an unknown?

As a manufacturer of transient voltage suppressors, we discarded the varistor a number of years ago because it will age, and aging increases the leakage current. The higher the current, the faster the aging — a runaway condition. Also the higher the ambient temperature, the sooner this can happen. We have marketed a 110-volt plug-in unit that is extremely effective, using a *Tranzorb** as the heart of a circuit; it is cost comparable to the circuits you describe. A comparison report of *Tranzorbs* versus various metal oxide varistors is available from General Semiconductor Industries.

Nanosecond ranges are becoming antiquated; picosecond ranges are the ones you must be concerned with.

Stephen J. Sorger,
Vice President
W. N. Phillips, Inc.
Lake City, Michigan

English, *si!* Metric, *no!*

Dear HR:

I have subscribed to *ham radio* almost since its inception, and have enjoyed it more than any of the other amateur magazines. When you added the metric units after the English measurements it was annoying, but I just cussed the confusion under my breath; when you made metric primary and English secondary, however, it was more than I could put up with. I refuse to have the metric system shoved down my throat.

S. D. Brokhausen, W5VMN
Georgetown, Texas

metrics made easy

Dear HR:

By choosing metrics for your primary system of measurement, you have again set the pace for the other amateur radio magazines. It's only through everyday usage that people will become conversant in the metric system; once they can speak *metric* as well as they do English, opposition to metric conversions will disappear.

David L. Campbell, W1CES
Boston, Massachusetts

**Tranzorb* is a registered trademark of General Semiconductor Industries, Inc., Post Office Box 3078, Tempe, Arizona 85281.

radiation hazard

Dear HR:

W1HR's editorial in the August issue should serve as a caution to the ham community to exercise great care in the operation of vhf and uhf installations. This is particularly timely in view of the accelerating trend toward these higher frequencies.

There is, however, another possible hazard area which seems to have been ignored by amateurs and commercials as well. This is the use of uhf/vhf handy-talkies. Here, too, their use is multiplying and our improving technology is compounding the problem. In the past, 1 watt was a high power for these units; now 5-watt units are common, and higher power is probably on the way.

Now it seems that this should be low enough to be negligible, but the inverse square law does apply, so that power densities in close proximity to the antenna can reach significant levels. In operation, the HT antenna is situated but a few inches from the eyes of the operator.

A quick calculation, based on an isotropic source radiating 1 watt, shows that the 5 mW/cm^2 occurs about 4 centimeters from the antenna. A 4-watt power would thus give this power level at 8 centimeters, which is about the distance that an HT would be located from its operator's eyes if he close-talks the unit — as most people do.

I recognize that the 5 milliwatt level is recognized as being safe by most authorities. Furthermore, we are protected by the fact that most HTs will not function for an extended period of time at the 5-watt level. But I do not believe that it would be prudent to continue to ignore this possible source of eye damage, and we certainly should limit HT power levels to no more than 5 watts.

Harley C. Gabrielson, K6DS
La Mesa, California

presstop

AUTOPATCHES WEREN'T BANNED by the FCC in its action on interconnects in late October, despite a rumor to that effect circulating on both coasts. What the Commission did was to reaffirm the requirement that a control operator be on duty any time a repeater is used for autopatch, incoming or outgoing. "Automatic control" of a repeater is not permitted during autopatch operation.

A Repeater During an autopatch is not a repeater, the Commission also decided, as it is not then operating as a repeater but as a conventional Amateur station making a phone patch. As a result, the rules governing regular Amateur station operation, not the repeater rules, apply during autopatches.

PHASE II OF FCC'S NEW CALLSIGN program is going into effect January 1, providing Advanced Class licensees with the opportunity to upgrade callsigns when renewing their licenses. After January 1, an Advanced Class licensee who wishes to exchange his present callsign for a 2x2 need only check item 13a on his renewal Form 610 and send it in during the 60 days prior to expiration. NOTE: Renewals received before January 1 will be processed without callsign change, so Advanced Class operators who wish to change callsigns should hold applications until after the first.

Requests For Callsigns other than "Group B" (2x2) will not be honored under Phase II, nor will Extras be permitted any choice other than a 2x1 after January 1 — under the present rules, an Extra can specify a new callsign from any group "higher" than his present, if he so wishes.

900 MHZ WILL BE CB'S NEW HOME, if, indeed, CB is to get a new home. At an open Commission meeting in October the Commissioners heard the FCC's Personal Radio Planning Group present three options: keep CB as is on 27 MHz, keep 27-MHz CB as is but add a new CB service at 220 MHz, or keep 27-MHz CB as is but add a new CB service at 900 MHz. They made no specific recommendations to the Commissioners, but discussed the three options at some length along with the factors affecting them (which definitely favored 900 MHz for expansion).

The Need For Any CB expansion was questioned by Commissioner Margita White and others, and after some discussion the Commissioners decided not to go ahead with a Notice of Proposed Rule Making. Instead, they had the Personal Radio Division staff prepare a Notice of Inquiry on the establishment of a new personal radio service in the 900-MHz region.

THE FCC'S FEE REFUND PROGRAM is supposed to get under way by the first of the year, with FCC licensees who paid more than $20 for their licenses first in line for reimbursement. Details of how all this is to be done have yet to be worked out, as the amount to be repaid is to be the fee less the license's "value" to the applicant (how much is a 2-letter call worth?). A Notice of Inquiry, soliciting comments on how such values can be derived, and other aspects of the program, will come out first. About all that seems certain at this time is that Amateurs who paid $25 for a special callsign are due to receive something back, and they'll be in the first group to be repaid.

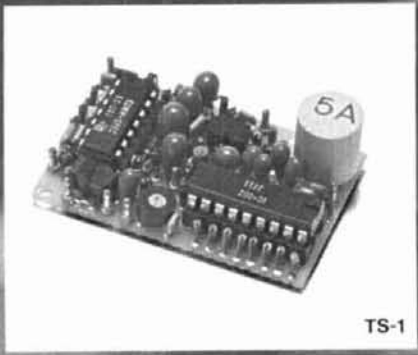
FCC Commissioner Margita White's term on the FCC has expired, and President Carter has nominated Ann P. Jones of Arlington, Massachusetts, to replace Mrs. White. Miss Jones has been general counsel to the Federal Home Loan Bank since last January. Mrs. White's departure from the Commission will be noted by the Amateur community. She's been particularly sympathetic to the needs of the Amateur service during her term.

LICENSE REVOCATIONS for bribing an FCC employee to issue unmerited 1x2 callsigns were handed out to two Ohio Amateurs in October. Revoked were the Amateur licenses of WA8ZDF/W8MZ and WB8AKU/WB8CPL, as an aftermath of the FBI's licensing investigation that resulted in the conviction of FCC Gettysburg employee Richard Ziegler in June, 1977. The revocations become effective in 50 days unless there is an appeal or the Commissioners choose to review the case.

THE NEGATIVE EFFECTS OF AMATEUR ANTENNA structures on neighborhood property values is the real issue in many anti-tower zoning laws and anti-antenna suits, the Personal Communications Foundation (PCF) reports. It's the issue in the N6QQ case, where the city of Cerritos has agreed that N6QQ's tower was structurally sound and its safety was not an issue.

Professional Help on the property value issue is being sought by the PCF for the N6QQ case, which the Foundation believes could have a negative effect on Amateurs' antenna rights nationally if it goes against N6QQ. Real estate appraisers, realtors, or others who could provide expert opinions on the antenna vs property value issue should contact Ken Widelitz, WA6PPZ, at the Personal Communications Foundation, 10960 Wilshire Boulevard, Suite 1504, Los Angeles, California 90024 (213) 478-1749.

BIOLOGICAL EFFECTS OF RF radiation will be studied by a newly formed professional group, the Biomagnetics Society. Formation of the group was announced at the International Microwave Symposium in Ottawa this summer, and its first president is Dr. Edward Alpen, professor of medical physics at the University of California at Berkeley. For membership information, write to Box 3651, Arlington, Virginia 22203.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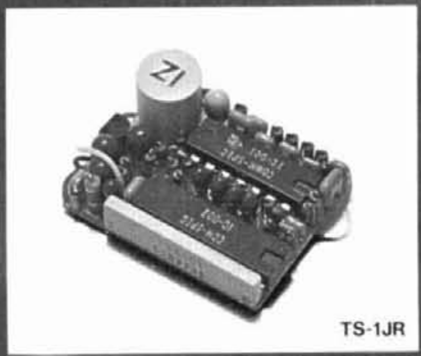
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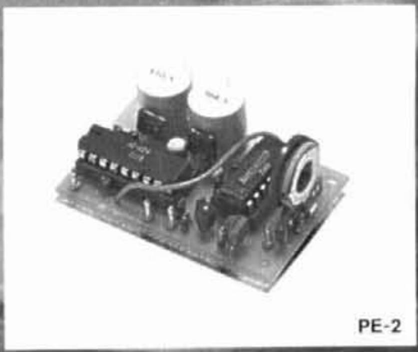
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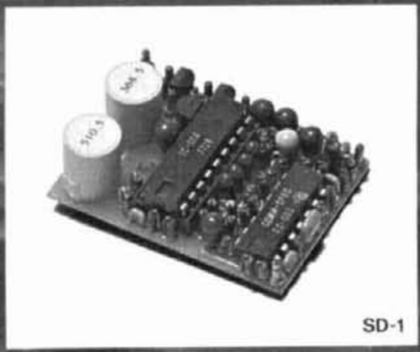
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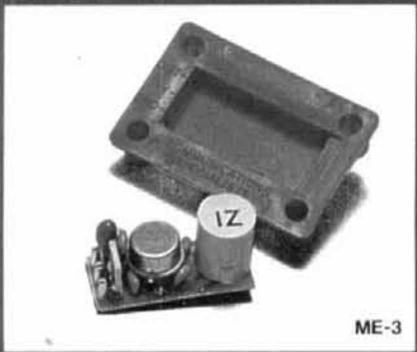

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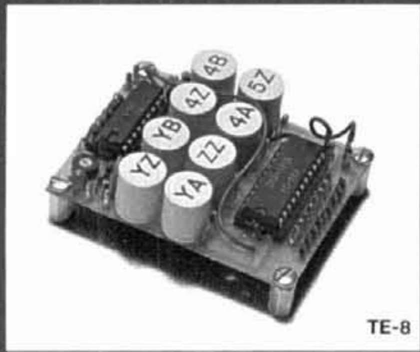

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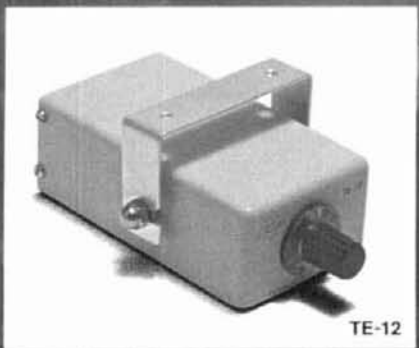

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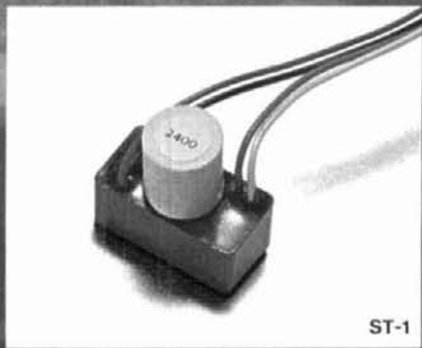

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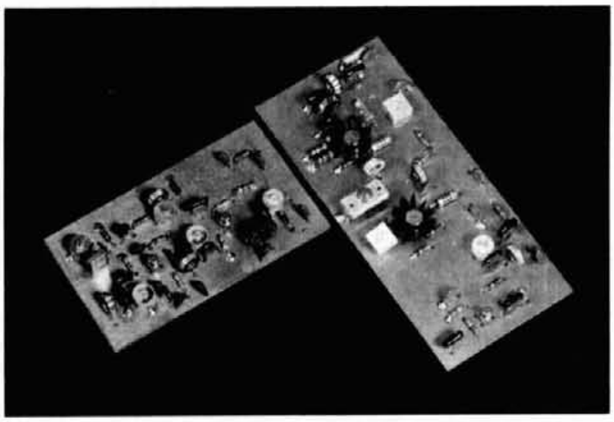
low-power transverter

for the high-frequency amateur bands

Is your QRP transceiver band limited? Try this simple transverter to obtain multiband operation

Low-power transceivers are becoming more popular each day on the amateur bands. Many of these rigs are single or dual band, usually for 40 and 20 meters. My home-brew 20-meter ssb transceiver fits this category. The need for 80-meter operation resulted in the transverter described here. A savings of both money and construction time can be realized if a transverter is used — rather than constructing an entire assembly for another band. The transverter eliminates the need for another ssb filter, VFO and associated components, and dial drive and indicator. The transverter can be used with any 20-meter transceiver or transmitter/receiver combination. Com-



Printed-circuit boards for the low-band transverter. Receiver board is to the left, transmitter board to the right.

By Mark Oman, WAØRBR, 427 East 11 St., Loveland, Colorado 80537

mercial QRP transceivers, such as the HW-7 and Ten-Tec PM series, can also be used with this transverter. Both ssb and CW modes can be used.

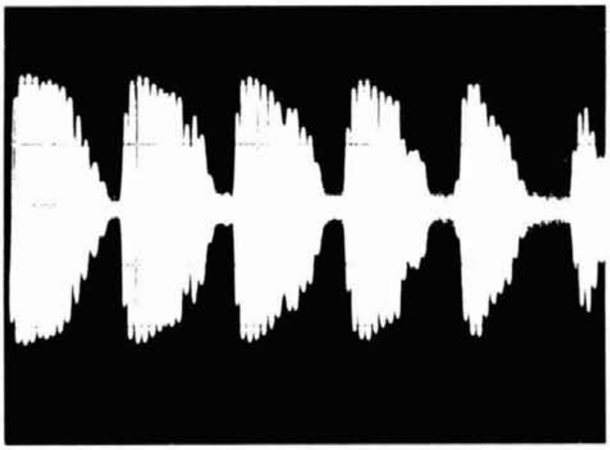
I built the transverter for 80 meters, but other bands can also be covered. With appropriate component changes for inductors and capacitors, the same PC-board layout can be used. **Table 1** lists values for these components for 160, 40, 15, and 10 meters. The transverter provides CW or ssb output of approximately 1-2 watts and needs a minimal amount of rf drive.

operation

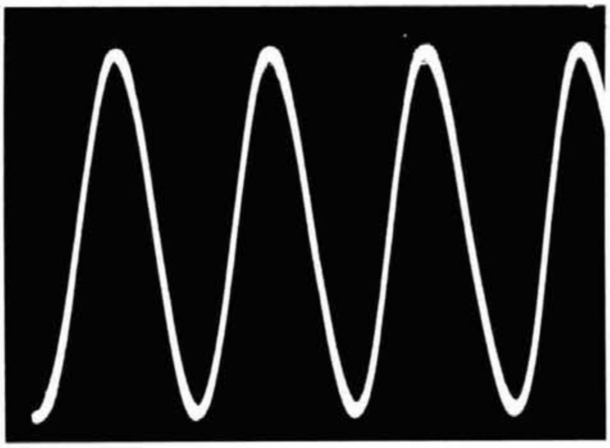
The basic principle of this transverter is to receive a 3.5-4 MHz signal, convert it to 14 MHz, and transmit a 3.5-4 MHz signal by converting 14 MHz to 3.5 MHz. A crystal oscillator of 18 MHz provides the signal to convert these frequencies. The oscillator operates on both transmit and receive modes.

receive mode

Referring to **fig. 1**, Q3 operates as an 18-MHz, third-overtone crystal oscillator with output tuned by L5 and C14. Q1 is a dual-gate mosfet rf amplifier tuned to the 80-meter band. The 80-meter signals are mixed with the 18-MHz oscillator output in Q2, resulting in a 14-MHz output across L4. A 50-ohm

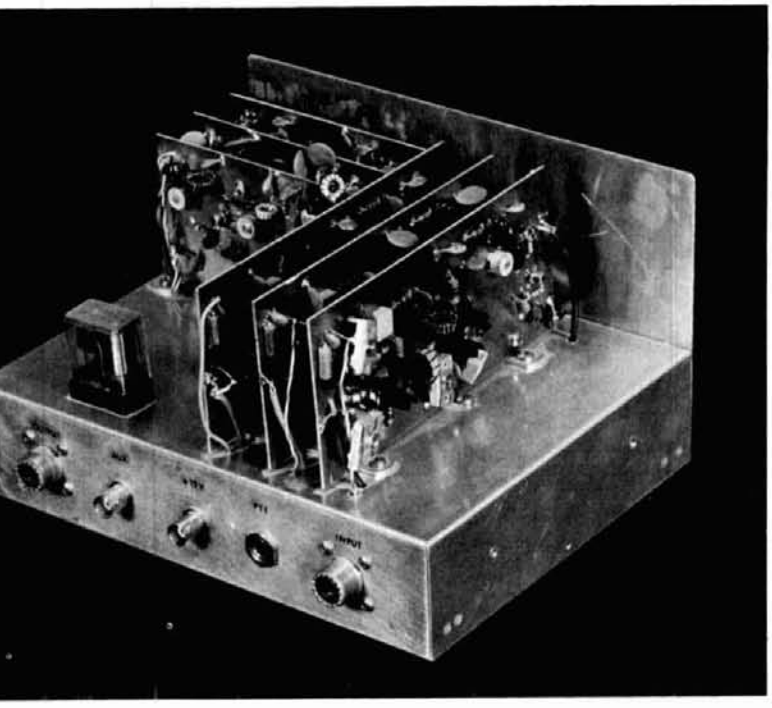


Transverter output on 80 meters showing ssb (upper) and CW (lower) modes as viewed on a Hewlett-Packard model 182 oscilloscope.



output impedance is obtained by using a capacitive divider across L4.

Rear view of the transverter showing the 10, 40, and 80 meter circuit board assemblies. The receiver converters are in the background, while the transmitter converters are in the foreground.



transmit mode

Q1 operates as a mixer, this time mixing 18 MHz with the 14-MHz input signal to produce a 4-MHz signal at Q2 base. Q2 is a class-A amplifier with about 100 mW output to drive Q3. A "T" network matches Q2 collector impedance to Q3 base impedance. Q3 is a linear amplifier. Its bias is adjusted by R13. L4, L5, C15, C16, and C17 are a pi network to match the Q3 collector impedance to 50 ohms. (This matching network also provides good harmonic attenuation.)

The transverter matching network design was based on the excellent articles in reference 1. Design was a lot easier than anticipated and circuit debugging was fairly easy.

Attenuator Z1 is a critical part of the circuit for transmitting clean signals. Most QRP rigs will need about 20 dB attenuation, as only 10-20 mW is needed to excite Q1. Too much drive will cause distortion on

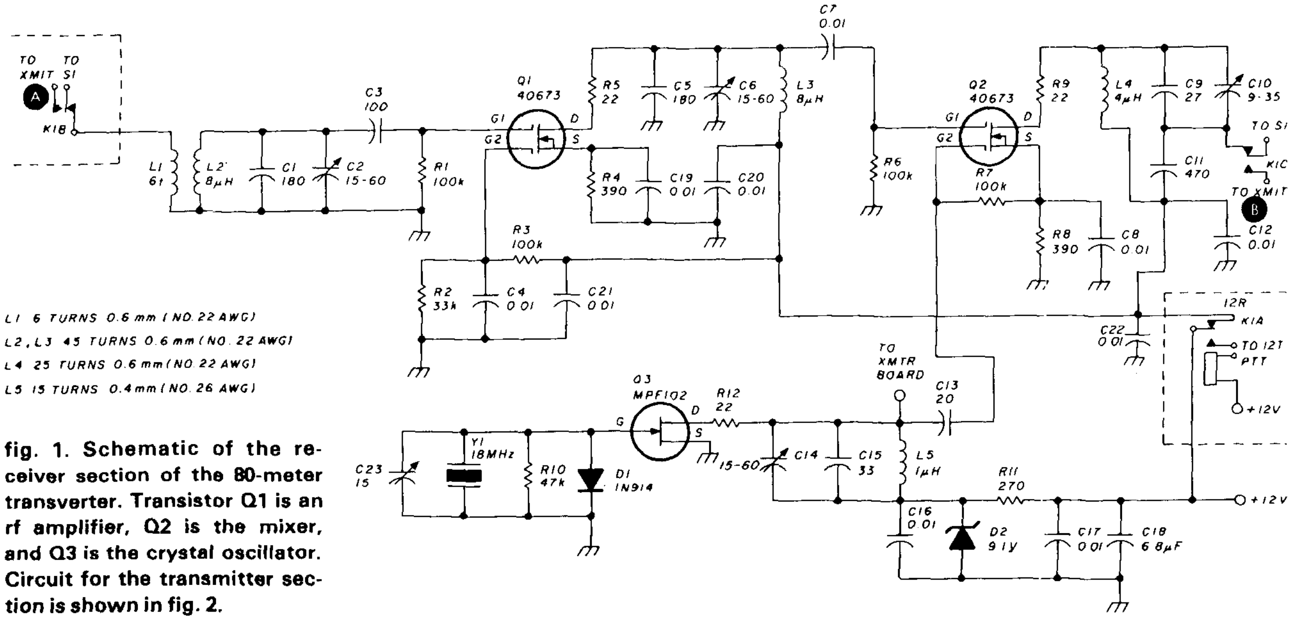


fig. 1. Schematic of the receiver section of the 80-meter transverter. Transistor Q1 is an rf amplifier, Q2 is the mixer, and Q3 is the crystal oscillator. Circuit for the transmitter section is shown in fig. 2.

ssb. Drive on CW, however, is not as critical. Some experimentation with Z1 may be necessary if you use ssb; *i.e.*, more or less attenuation may be needed with your rig. **Fig. 2** gives component values for the attenuation needed. The attenuator provides the additional bonus of a 50-ohm load to the driving transmitter, a near-must with solid-state rigs.

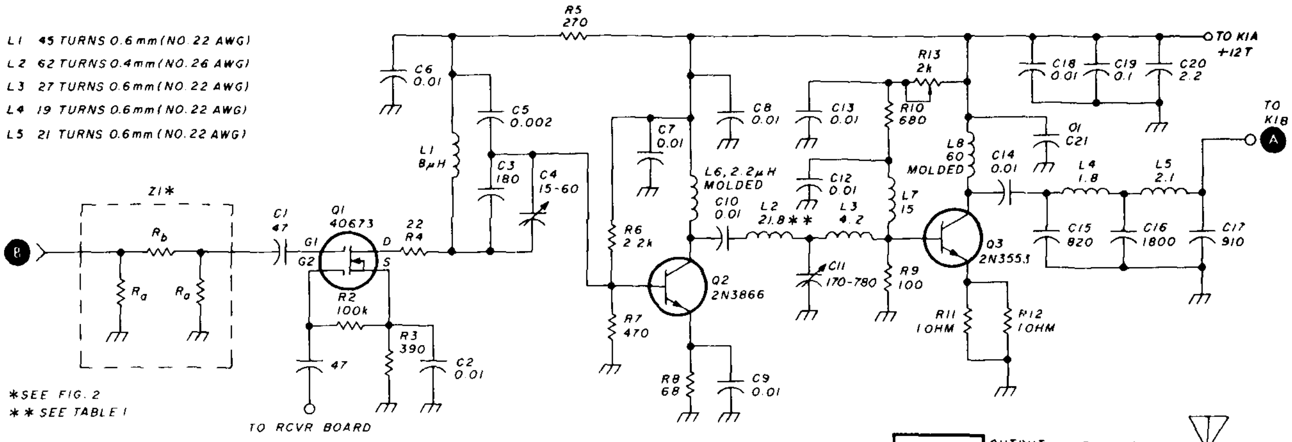
Liberal use of bypass capacitors on the dc lines helps to prevent oscillation in both transmit and receive modes.

construction

Transverter construction is easy because of PC-board layout and use. The receiver and transmitter sections are built onto separate boards. Circuit-board patterns for both the transmit and the receive board are shown. Double-sided, glass-epoxy board is used, with one side acting as a continuous ground plane. The copper around the component leads is drilled out using a large drill bit.

PC boards are definitely the answer to home construction. They provide clean layout, ease of construction, and, almost always, superior operation. Board layout is easy and provides a convenient way to make changes and experiment with your own ideas.

The two boards are wired together using RG-174/U cable. I mounted mine in an enclosure where



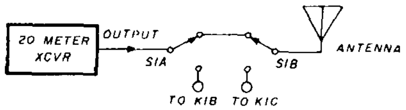


fig. 2. Circuit of the *transmitting section* of the 80-meter transverter. Transistor Q1 is the mixer with oscillator injection from the receiver crystal oscillator; Q2 is a class-A driver, and Q3 is a linear power amplifier. Some adjustment of the values of the input attenuator Z1 may be required for correct operation with different rigs.

table 1. Values for transverter components for 160, 40, 15, and 10 meters.

receiver assembly

band	L1	L2(μH)	L3(μH)	C1, C5 (pF)	C2, C6 (pF)
160m	7t	16 53t/T68-2	8 45t/T68-2	430	15-60
40m	4t	4.3 29t/T50-2	4.3 29t/T50-2	82	15-60
15m	3t	1.5 22t/T37-6	1.5 22t/T37-6	18	9-35
10m	3t	1.1 19t/T37-6	1.1 19t/T37-6	5	9-35

band	C14(pF)	C15(pF)	L5(μH)	Y1(MHz)	bypass capacitor (μF)
160m	15-60	33	1.2	16	0.1
40m	15-60	15	1	21.5	0.005
15m	15-60	—	0.5	35.5	0.001
10m	9-35	—	0.5	42.3 42.8	0.001

transmitter assembly

band	L1	L2	L3	L4	L5(μH)
160	16 53t/T68-2	44 90t/T68-2	8.5 40t/T68-2	3.6 25t/T68-2	4.2 27t/T68-2
40	4.3 28t/T68-2	11.8 45t/T68-2	2.25 21t/T50-2	1 14t/T50-2	1.1 15t/T50-2
15	1.5 22t/T37-6	4 32t/T50-6	0.8 14t/T50-6	0.33 10t/T37-6	0.37 11t/T37-6
10	1.1 19t/T37-6	3 27t/T50-6	0.6 14t/T37-6	0.25 9t/T37-6	0.28 10t/T37-6

band	C3	C4	C5	C11	C15	C16	C17(pF)
160	430	15-60	5000	170-780*	1800	4000	2000
40	82	15-60	1000	50-380	430	1000	500
15	18	9-35	390	37-250	150	330	180
10	5	9-35	300	16-150	110	240	130

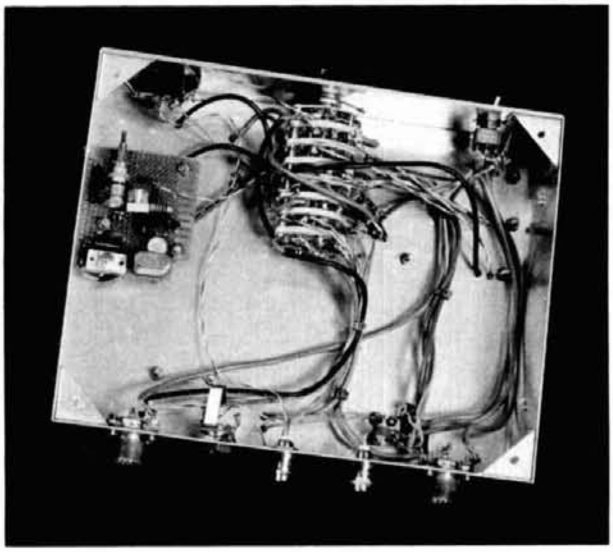
band	L6(μH)	L7(μH)	L8(μH)	bypass capacitors (μF)
160	4.7	33	150	0.1
40	1.5	8	33	0.005
15	0.47	2.7	12	0.001
10	0.27	2.2	10	0.001

*In parallel with 500

they are grouped with boards for 40, 15, and 10 meters to make an all-band transverter. A 3PDT relay switches +12V, the 20-meter transceiver, and the antenna between transmit and receive modes. S1 allows either the transverter to be used or the main rig to be operated straight through. Heat sinks should be used at Q2 and Q3.

alignment

The first step in aligning the transverter is to obtain

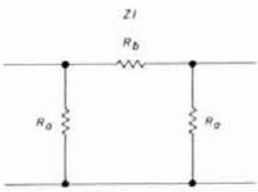


View of the chassis showing the bandswitch and internal wiring. The small board on the left is a WWV receiver converter.

18-MHz operation from Q3. Adjust C14 for maximum output using either an rf probe or a communications receiver. C23 can be used to adjust Y1 to exactly 18 MHz. The next step is to adjust C2, C6, and C10 for maximum signal strength while receiving an 80-meter signal.

Before applying drive on transmit, adjust R13 for minimum Q3 collector current. Next apply a low-level signal to Q1 (approximately 8V p-p at G1 of Q1). With a 50-ohm dummy load connected to the output, adjust C4 and C11 for maximum output.

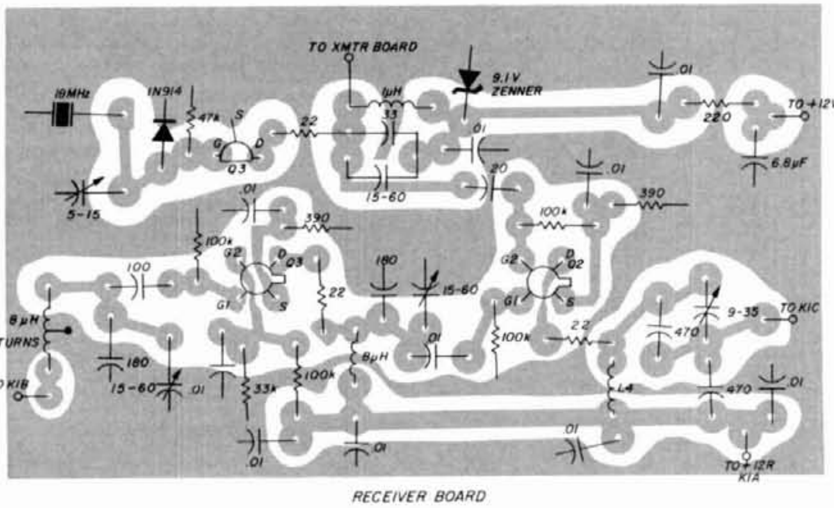
Monitor the signal on a separate 80-meter receiver for best tone characteristics. For ssb adjust R13 for a collector current of 5 mA. Adjust C4 and C11 for maximum output. Monitor the ssb signal on an 80-



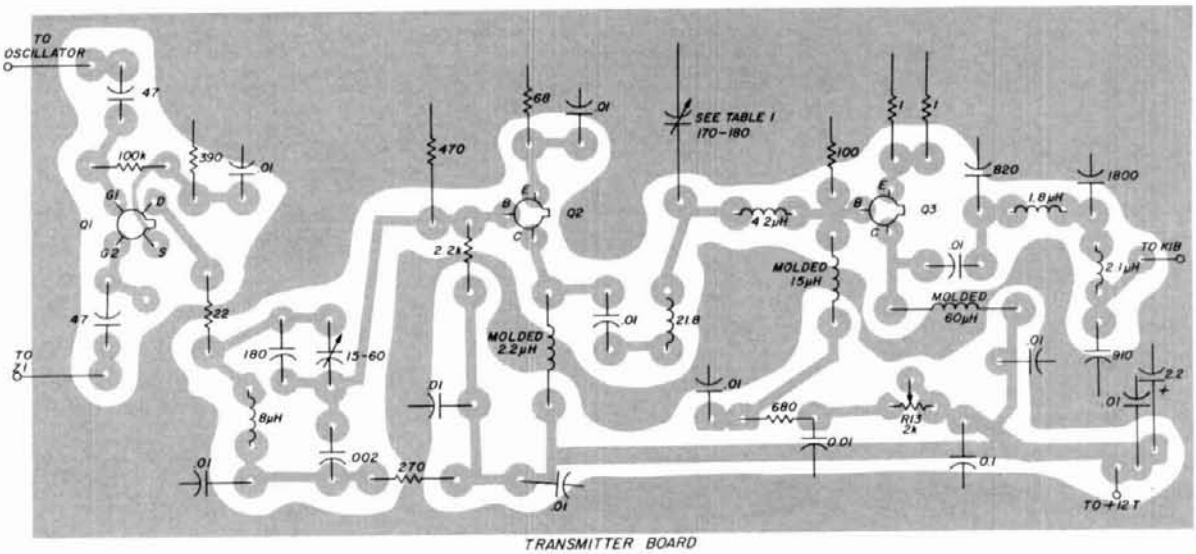
attenuation (dB)	R_a(ohms)	R_b(ohms)
3	292	18
6	150	37
9	105	62
10	96	71
12	83	93
15	72	136
20	61	248
25	56	443
30	53	789

fig. 3. Component values for different attenuation ratings of the pi attenuator, Z1, in fig. 2.

fig. 4. Component layouts for the printed circuit boards used in the low-band transverter.* Receiver board is shown on top, transmitter board at bottom.



RECEIVER BOARD



TRANSMITTER BOARD

meter receiver and readjust drive for no distortion (see photos). Readjust C4 and C11 for the best-sounding ssb signal. This completes alignment. The transverter should now be ready for on-the-air use.

operation

Because of the transistor used and the output network, the transverter *must* be operated into a 50-ohm load. Any departure from 50 ohms can cause the final stage to self-oscillate. Adding a 40V zener from Q3 collector to ground would be helpful in preventing damage to Q3 should oscillations occur. A transmatch is a great help in providing a nonreactive load for the transmitter.

With the output displayed on an HP-182 100-MHz scope, no harmonic energy could be observed. The CW and ssb outputs are shown in the photos.

Approximately two watts output was obtained on CW and one watt on ssb. The networks are sufficiently broadband to allow operation across the 80-meter band.

in conclusion

The transverter has provided good 80-meter coverage. Contacts have been made between the East and West Coast and into Mexico. Because QRP is my main interest, the one-watt output level is quite adequate. Under poor band conditions a linear amplifier can be added to increase output level.

*Full-size printed-circuit layouts are available from *ham radio*; please send self-addressed, stamped envelope with all requests.

reference

1. DeMaw and Rusgrove, "Learning to Work with Semiconductors," *QST*, April - September, 1975.

ham radio

lightning protection

for the amateur station

A lightning strike can be devastating. This article provides a thorough treatment of how to avoid catastrophe in the ham shack

Lightning protection is a subject that should be of interest to every amateur and especially to those who have high antennas. It's a subject that's had little coverage previously in amateur publications despite the fact that there are some very effective protection techniques, which are widely known and applied to professional communications installations. Many of these techniques can be applied to the typical amateur station with minimal expense and effort. Some techniques, which involve more effort and expense, might be warranted only in situations where the probability of being struck by lightning is high.

Before protection techniques are discussed, it's desirable to have an understanding of exactly what causes lightning, the magnitude of a lightning stroke, and the probability of being struck.

The Earth's atmosphere is constantly being bombarded by cosmic rays. When these rays collide with gas molecules electrons are separated from some molecules, creating positive ions. Similarly, some molecules capture a free electron and become negative ions. It's estimated that our atmosphere contains about 4×10^3 ions per cm^3 (6.5×10^4 per $inch^3$). At altitudes above 64 km (40 miles), the number of ions exceeds the number of neutral molecules; this is the region known as the ionosphere. At low altitudes, this number is insignificant compared with the number of neutral molecules. Still, the presence of these ions makes it possible for the air to conduct electricity to a small degree.[1]

Atmospheric makeup. The atmosphere taken as a whole has a net positive charge of about 10^6 coulombs. The Earth's surface has an equal negative charge, and a potential difference of about 3×10^5 volts exists between it and the electrosphere. The electrosphere is the region beginning at about 48 km (30 miles) up, in which the resistivity is sufficiently low so that there's no significant voltage gradient. (The ionosphere has the still lower resistivity necessary to reflect radio waves.) In fine weather, a constant flow of electrons occurs from the earth to the electrosphere, resulting in an electric field of about 100 volts per meter near ground. Lightning discharges return these electrons to earth at a rate sufficient to sustain a balanced dynamic system globally.[2,3]

For lightning to occur, a localized region of the atmosphere must attain sufficient electrical charge to produce a breakdown of the air molecules. The electric field near ground level rises to 500 volts per centimeter below a developing thundercloud and much higher still when lightning is about to strike.

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The inside of a thundercloud is a turmoil of water, ice, and dust particles together with strong wind currents and temperature gradients. Although the mechanism is not totally understood, the result of this turmoil is a concentration of positively charged particles rising to the top of the cloud; negatively charged particles are concentrated in the lower areas of the cloud. This negatively charged region at the base of the cloud repels free electrons on the ground, resulting in an area beneath the cloud that's positively charged both with respect to the cloud and with respect to surrounding earth. Conditions now are correct for lightning to strike.

Evolution of a lightning strike. A lightning flash begins with a virtually invisible stepped leader, which travels down from the cloud toward the ground. Each step covers a distance of about 46 meters (150 feet) in less than one microsecond; the time between steps is about 50 microseconds. As this stepped leader progresses, it ionizes the air through which it passes, making it a good conductor. When the leader reaches within a few hundred meters of ground, ionized streamers begin to rise from the ground to meet it. Then the conductive path from cloud to ground is complete, and the visible portion of the bolt, known as the return stroke, begins.

It is the *return* stroke that has the destructive effects against which protection is needed. As soon as the conductive path is completed, electrons start flowing rapidly to ground. This action starts at the point of contact between the stepped leader and the rising streamer, and the greatly increased current causes the ionized path to glow brightly and get very hot — up to 2×10^4C (6.5×10^4F). The region of high current and brightness moves upward to the cloud at a speed of over 9.6×10^4 km (6×10^4 miles) per second, drawing electrons from higher and higher in the cloud. This contrasts with the speed of the stepped leader, which typically averages only 384 km (240 miles) per second. Although the region of high-current density and hence the visible flash moves upward, the actual flow of electrons is downward.

After the first return stroke usually enough charge remains in the cloud to initiate a second leader. This usually occurs within 70 milliseconds or less; because of remaining ionization, this leader usually follows the path of the previous stroke and travels directly to earth in one step of about one millisecond. For this reason it's called a *dart leader*, and, like the stepped leader, it's followed by a return stroke. Although the average bolt flashes only twice, about ten per cent will have as many as ten flashes; occasionally bolts will have up to twenty flashes over a period of about one second.

Electrical analysis. For all practical purposes a lightning bolt may be considered a constant-current source. That is to say, when it strikes an object protruding above ground, the current that flows through that object to ground will be the same, regardless of whether the object is a metal tower with a resistance of 1 ohm or a tree with a resistance of several hundred ohms. This is because the atmosphere's resistance is so high that the series resistance added to the path by the object struck is insignificant. The resistance of the object struck becomes important when the voltage drop across the object and the power dissipated in the object are considered.[4]

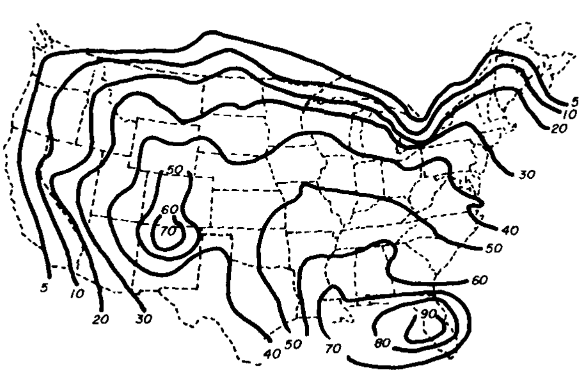


fig. 1. Map showing average annual number of thunderstorm days in the continental United States.

The magnitude and duration of the current in a lightning bolt will vary from stroke to stroke. The stroke-current waveform consists of a rapid rise to the peak current, followed by a more gradual decay. A waveform of this type is described with two numbers: the rise time to the peak value and the decay time to 50 per cent of the peak value. A typical lightning bolt would have a rise time of about 2 microseconds and a decay time of about 40 microseconds. The peak current in a stroke will exceed 1.7×10^4 amps in 50 per cent of all strokes. It will exceed 6×10^4 amps in 10 per cent of all strokes, 1.2×10^5 amps in 1 per cent of all strokes, and 2.4×10^5 amps in 0.01 per cent of all strokes.[5]

the probability of lightning damage in the U.S.

The probability that any given object will be struck by lightning depends on two things: the frequency and type of thunderstorms at the location of the object, and the object's height above average terrain. Observations made over a period of years have resulted in the compilation of an accurate map showing the average number of thunderstorm days per year in the United States (**fig. 1**). Note that this map does not take into account the possible occurrence of more than one thunderstorm in an area on any given day. Therefore it's somewhat conservative as an indicator of the total number of thunderstorms likely to occur in a year, particularly in areas having larger numbers of thunderstorm days.

Thunderstorms are classified either as convection storms or frontal storms. Convection storms, which account for the majority of thunderstorms, are local in extent and relatively short in duration. Frontal storms extend over greater areas and may last for several hours. Statistical data has been tabulated to predict the expected number of lightning strokes to ground per square mile (assuming flat terrain) per thunderstorm day. This number is called the *stroke factor.* The stroke factor for convection storms is 0.28; for frontal storms it is 0.37.[6]

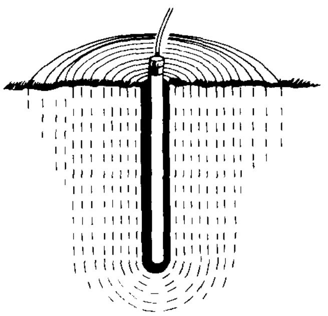


fig. 2. The concentric shell concept as an aid to understanding the resistance of the earth surrounding a ground rod.

In areas of flat terrain, the probability of lightning striking any given point is extremely low, but an object protruding above ground will attract lightning that would have otherwise struck all other points within a circular area surrounding the object. The radius of this area depends on both the height of the object and the intensity of the stroke. An object will attract a 2×10^4-amp stroke that would have struck anywhere within a radius of twice the height of the object had the object not been present. This radius increases to six times the height of the object for a 6×10^4-amp stroke and ten times the height of the object for a 1.35×10^5-amp stroke.[7]

the importance of proper grounding

Three distinct situations are associated with a lightning strike that can result in damage to electronic equipment. The most obvious and potentially destructive is the effect of a direct strike. Less obvious, but still capable of causing considerable damage, are two secondary effects: induced voltage and ground current.

A lightning bolt is surrounded by an intense magnetic field. The bolt itself may go harmlessly to ground, but its magnetic field will induce a voltage transient into any wires it encounters, such as antennas and feedlines, rotor-control cables, power lines, or telephone lines. When a lightning bolt goes to ground, current flows through the ground in all directions away from the point of the strike to dissipate the sudden excess accumulation of charge. This ground current can cause a significant potential difference to exist between different ground points. Improperly grounded equipment can end up in the path of some of this ground current. The lightning protection techniques to be described can prevent equipment damage from all three of these effects.

protecting your gear

Lightning protection comes in increments. An unprotected station will be wiped out by any kind of strike. A partially protected station will incur less damage and may completely escape damage from a small bolt. A carefully planned and executed lightning protection system will protect against just about anything.

There are three basic protection techniques. Everything described here will contribute to one or more of them. The first technique is to send as much as possible of the lightning stroke directly down your tower and into the ground. The second technique is to make it as hard as possible for any of the energy of the strike to get to the equipment. The third and final technique is to control the path of any energy that reaches the equipment so that it finds its way harmlessly to ground.

Grounding systems. The first technique, that of sending as much of the current as possible directly to ground, requires grounding your tower. This isn't always easily accomplished. There's more to it than just driving a rod into the ground and connecting it to the tower. Any ground system has a resistance, which can be measured. This resistance determines the voltage level to which the tower will rise when struck by lightning. The magnitude of this resistance depends on certain characteristics of the soil that determine its resistivity; *i.e.*, composition, temperature, moisture content, and salt content.

A ground rod driven into soil of uniform resistivity radiates current in all directions. To understand the resistance of the soil surrounding the ground rod, think of the rod as being surrounded by an infinite number of thin concentric shells of soil all of equal thickness (**fig. 2**). The greatest resistance is in the shell directly next to the ground rod, because it has the smallest cross-sectional area at right angles to the current flow. Each succeeding shell is larger in cross section and therefore has less resistance. At a small distance from the rod the area of each shell is so large that its resistance is negligible compared to the resistance of the shell directly next to the ground rod. The resistance varies inversely with the cross-sectional area. Measurements have shown that 90 per cent of the total resistance surrounding a ground rod is usually within a radius between 1.8-3 meters (6-10 feet) of the rod.

Soil considerations. The composition of the soil will depend on your location. Whatever soil you have you'll have to live with. Clay and loam are the most desirable soils for low resistance. Sand or gravel

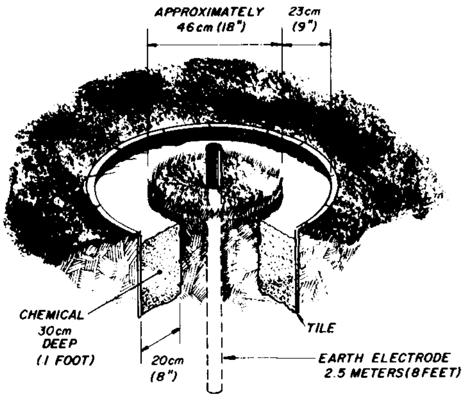


fig. 3. Method of reducing resistance of a ground connection by chemical means. The tile may be eliminated with little effect.

increase the resistivity in relation to their proportions, and soil that is mostly sand or gravel has fairly high resistance.

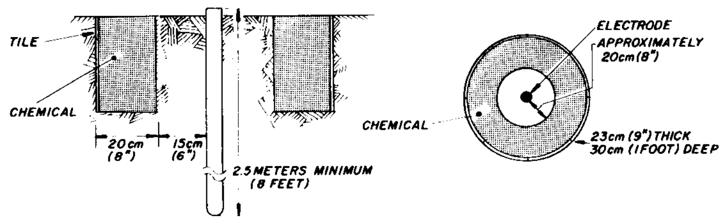
Whatever type of soil you have, a deep rather than a shallow ground is better for several reasons. Soil is seldom of uniform resistivity at different depths. The soil near the surface generally has higher resistivity than that at deeper levels because of wetting and drying out with seasonal variations. Deeper soil is more stable and less subject to such variations. Usually it has a higher moisture content than surface layers.

As the soil temperature decreases, its resistivity goes up. Frozen ground is a very poor conductor, as the negative temperature coefficient of soil resistivity increases sharply below freezing. Any ground rod should, at a minimum, extend several feet below the frost line in your area. Otherwise an early spring thunderstorm may occur while the ground is still frozen, and your ground system will be greatly reduced in effectiveness. A 2.5-meter (8-foot) ground rod is the absolute *minimum* that should be used.

reducing soil resistance

Soil resistivity may be reduced artificially by increasing its salt content. A doughnut-shaped trench can be dug surrounding the ground rod. Alternatively, a length of drain tile can be buried next to it. The trench or tile is filled with a chemical such as magnesium sulphate, copper sulphate, or ordinary rock salt. Rain and snow will dissolve the chemicals and wash them into the soil (**fig. 3**).

Deep-driven grounds are becoming the most popular and economical method for obtaining low-resistance ground connections. For ease of handling and driving, sectionalized *Copperweld* ground rods may be used. These rods are available in several diameters



and lengths. They are threaded on both ends, and special threaded couplings are used to join them. A driving bolt is attached to the top of the section being driven to protect the threads from damage (**fig. 4**).

Another effective method of reducing the resistance of a ground connection is to parallel several ground rods. For this technique to be effective, the rods should be between 1.5-3 meters (5-10 feet) apart. With less than 1.5-meter (5-foot) spacing, the conducting paths from adjacent rods will overlap excessively. With more than 3 meters (10 feet) of spacing, the reactance of the connecting wires becomes a detriment.

If multiple ground rods are used around the tower base, each rod should be securely fastened to the tower base, and also to a buried ring of 6.5 mm (no. 2 AWG) copper wire surrounding it. Guyed towers should have at least one ground rod at each guy anchor point; these rods should also be tied back to the ground ring at the tower base. In areas of high soil resistivity, buried radials spaced every 60 degrees and up to 46 meters (150 feet) in length, if possible, may be connected to the ground ring to lower the ground system resistance.

There's no hard and fast rule as to how low the ground system resistance should be for good protection. A value considered acceptable by some electric power systems and protection codes is 10 ohms, so this is a good objective to shoot for. Where the soil is mostly clay or loam this value is achievable, but in areas with rocky or sandy soil it may be economically unreasonable to attain.

Accurate measurement of ground resistance requires the use of sophisticated equipment not usually found in the amateur station. An ohmmeter is totally useless due to voltage gradients present in the soil from stray power-system currents and electrochemical effects. However a method that can be used with reasonably good success requires only

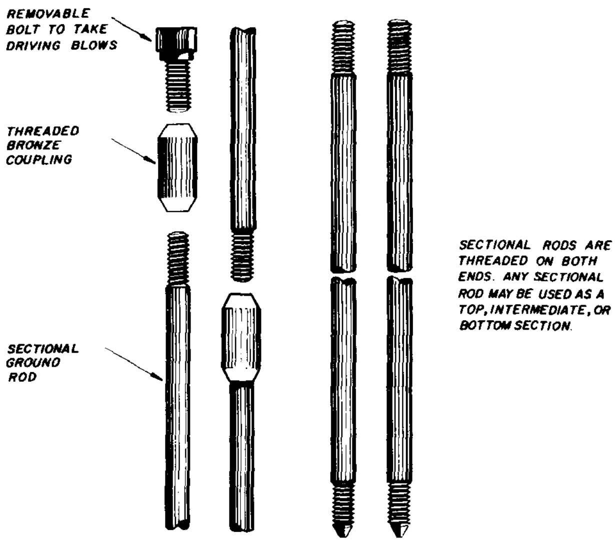


fig. 4. Copperweld sectionalized ground rods.

equipment that many hams would have or could borrow. It involves using an isolation transformer and a *Variac* (autotransformer) to pass 60-Hz ac through the ground between the ground rod to be tested and a known good ground, such as the power system neutral line. The voltage and current are measured, and the resistance is computed from Ohm's law. There's a very definite shock hazard associated with this method, so be careful! See **fig. 5**.

A question is often asked as to how large a ground conductor is necessary in a lightning-protection system. Often wire much heavier than necessary is used. A lightning stroke is of very short duration, and the heat produced in the wire by the current is limited. At a minimum, the wire must be large enough so that it's not heated to its melting point. A 2.6-mm (no. 10 AWG) copper wire can be sufficient to withstand a lightning stroke with 2.5×10^5 amps peak current and a 40-microsecond decay time. This amplitude is exceeded by only one stroke in 10^4. A 4.1-mm (no. 6 AWG) copper wire is recommended to provide an adequate safety factor in above-ground applications. For buried applications, 6.5 mm (no. 2 AWG) copper wire is recommended because of corrosion. Aluminum wire should have a cross-sectional area 1.6 times as large as the recommended copper gauges for the same current-handling capacity.

tower and rotator considerations

With the tower securely grounded at the base, the next step is to ensure that this good solid ground extends all the way to the top. Although you'd expect good continuity between sections in a fixed tower, the joints are designed for mechanical strength only, and the possibility always exists that oxidation and corrosion will cause increased resistance between sections. A good practice is to bridge each section joint on each leg with a short length of 4.1-mm (no. 6 AWG) copper wire secured with ground clamps. This is particularly important at the hinge joint of foldover towers.

Crankup towers are harder to deal with, which is unfortunate because they're also more likely to be a less-than-ideal path to ground for lightning — unless grounding straps are added. If the tower is lowered only occasionally for antenna work, the same techniques as those for a fixed tower can be used. The obvious inconvenience is that you'll have to climb the tower (make sure all safety stops are properly engaged first) and remove the grounding straps before the tower can be lowered. If the tower is raised and lowered frequently, about the only thing you can do is run grounding straps from the top of the top section all the way to the top of the bottom section. The straps should be pulled straight when the tower is fully raised, and of course they will hang down when it is lowered.

To protect your rotor, one or more bonding straps should be connected between the mast above the rotor and the tower legs. Leave no more slack than necessary to allow full rotation. Stranded wire, such as automotive battery cable, will stand up better to repeated flexing than will heavy solid wire.

The mast itself, or an air terminal clamped to the top of it, should extend far enough above the antenna so that its cone of protection includes the antenna and protects it from a direct hit. This mast extension should be at least half the turning radius of the antenna. The elements of an antenna are usually heavy enough to escape damage even if hit directly, and most antennas are at dc ground potential; but the path that provides this dc ground may not withstand the stroke current. Items such as ferrite baluns, traps, gamma matches, and element insulators are vulnerable and hard to protect in case of a direct hit on the antenna itself.

If the topmost point on the antenna system is the vertical element of a ground plane, make sure it is dc grounded at the base with an rf choke. This choke should have the lowest inductance possible for the frequency range of the antenna and should be made of the heaviest gauge wire practical.

If the tower is a wooden pole or other nonconducting support, it should be topped with an air terminal connected to the ground system with a 4.1-mm (no. 6 AWG) copper wire stapled to the pole on the side opposite the antenna feedline. A roof-mounted antenna system should be grounded similarly.

The shields of all coaxial feedlines should be bonded to the tower at the top and bottom, and at 15-meter (50-foot) intervals on taller towers. This is easily accomplished by using bulkhead coax connectors mounted on plates, bolted to the tower at the appropriate levels. Be sure to use waterproof connectors or take other suitable steps to keep moisture out of the coax.

residual lightning energy

All the techniques covered to this point have been concerned with shunting as much of the direct lightning energy as possible directly to ground. Now the emphasis will shift to the leftover energy that has found its way onto feedlines, rotor cables, etc., and to techniques for keeping it out of the rig. It all comes down to making these conductors look like a high impedance to the lightning. Remember that a lightning bolt is a series of pulses having rise times on the order of 2 microseconds. This makes it similar to a 500-kHz rf signal in terms of the effects of reactive elements it may encounter. Specifically, it's desirable to introduce *series inductance* in any conductors leading from the tower to the rig.

Obviously it's impractical to break the run of coax and place coils in series with the inner and outer conductors. While these might stop the lightning coming down the line, they would just as certainly also stop the flow of rf energy from the transmitter to the antenna. If the antenna is at dc ground as recommended earlier, the same potential will be impressed onto both the inner and outer coax conductors by a lightning strike. At every convenient opportunity, the coax should be bent at as sharp an angle as allowed by its minimum bending radius. The lightning pulse wants to go in as straight a line as possible, so these bends appear to it as an increased impedance. As far as an rf signal is concerned, as long as the dimensional relationship between coax inner and outer conductors is maintained, it makes no difference how often it is bent. Any excess feedline should be wound into a coil before entering the building, as this further increases the impedance it presents to lightning. Rotor cables and any other lines from the tower to the station should be treated similarly.

An additional technique that may be employed is to run the feedlines and control cables through a length of metal conduit. The magnetic field generated by lightning-stroke currents will perceive the conduit as one big shorted turn, resulting in another desirable increase in series impedance. This is particularly effective if the cables are run through conduit for 6 meters (20 feet) or more.

I'd like to emphasize that the isolation techniques just described are effective *only* if a good ground has been established at the tower. All the current is going to go to ground somewhere, and a low-resistance tower ground, together with a high impedance from the tower to the rig, only establishes how the current is divided up between these two paths.

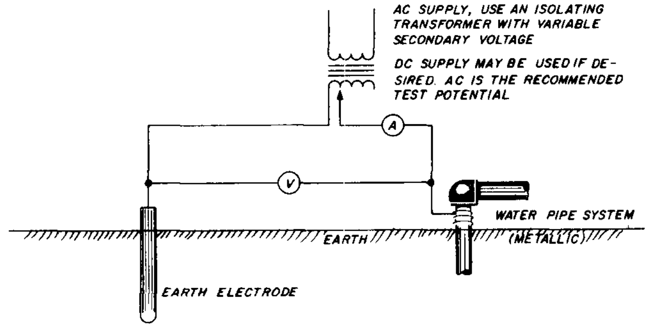


fig. 5. A simple method of measuring ground resistance.

station grounds

At this point, the protection job is two-thirds done. With good tower grounding and rig isolation, most of the stroke current will go straight down the tower to ground. The remaining task is to control the ground path followed by the remaining current which comes into the rig so that no equipment damage results.

It would be nice if it were possible to arrange things so that no ground path at all goes through the rig. In fact it's possible but seldom practical to do exactly that. It's worthwhile to examine what it would take to accomplish this, if for no other reason than to help in understanding the protection techniques to be followed in the more practical alternatives.

Let's assume that the only connection from the tower to the rig is a single feedline. Let's also assume that you have no other connection between the rig and the rest of the world; *i.e.*. no ac power lines, no telephone, nothing. The rig must be battery powered or have its own ac generator. The final assumption is that the entire rig, including its power source, is perfectly insulated from ground; no leakage, and sufficient clearance such that no arc-over to ground will occur.

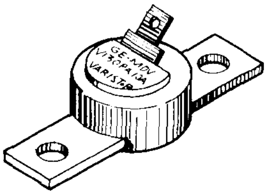
Now if lightning strikes the tower, let's see what happens. If it is an "average" stroke of 1.7×10^4 peak amps, and if the tower ground resistance is 10 ohms, the tower, feedline, and rig will all rise to 1.7×10^5 volts above ground. This number is conservative because it ignores the effects of inductance in the grounding path. In many cases a significant inductive reactance will occur in series with the resistance, and the potential rise will be higher for a given stroke current. If the assumption is good that the rig is perfectly insulated from ground no current will flow, and no damage will occur.

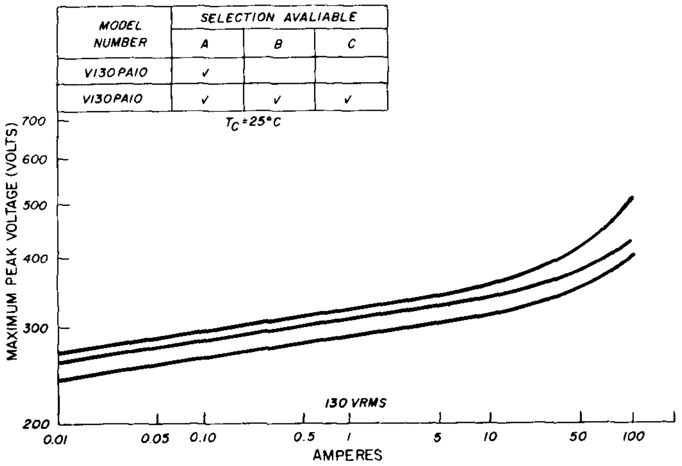
practical considerations

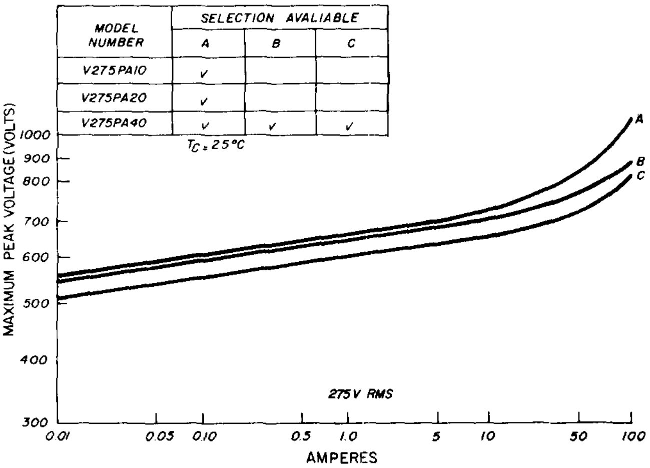
It should be obvious at this point that this kind of situation is seldom likely to exist. In the real world, the rig is not perfectly insulated from ground and connections exist to power lines and to the telephone ground bus. This morass of ground interconnections may be fine in terms of signal distribution and ac safety, but it can spell disaster when lightning strikes.

Consider where the current is actually going to flow. For example, it may come down the coax-feedline shield to the linear amplifier chassis, across this chassis to the input connector, along the input coax shield to the exciter, up the power cable ground wire to the power supply, through the power supply printed-circuit board to the ac line cord, and down the *third wire to the ac conduit.* A long and devious route such as this will look highly inductive to the lightning pulse. As a result significant voltage drops will occur along the way with a high probability of equipment damage.

fig. 6. Ratings for a high-power zinc-oxide varistor. Note the sharply rising voltage characteristic above 10 amperes on parts rated for up to 2750 amperes peak current.







model number	rms applied voltage 50-60 Hz volts	recurrent peak applied voltage volts	dc applied voltage volts	energy joules	average power dissipation watts	peak current amperes	characteristics: varistor peak voltage at 1MA ac min. volts	max. volts	thermal resistance hot spot to case °C/watt
V130PA10				10	8	1200			6.8
20	130	184	170	20	15	2750	185	255	3.6
V275PA10				10	4	600			13.7
20	275	389	360	20	7	1200	390	523	7.8
40				40	13	2750			4.2

system. This brings us back to that small portion of the stroke current that will go to ground through whatever path it can find at the rig. It can follow two routes, and each requires attention if damage is to be prevented.

The first of these routes encompasses all direct grounding made at the rig. This grounding can include the shield of signal-carrying leads between different pieces of equipment, the third wire in ac line cords, and wires connected from the ground screws on various pieces of equipment to a main-station

Single-point grounds. The solution to this problem is single-point grounding of each piece of gear together with bonding between each item. Single-point grounding prevents ground current from flowing through a piece of equipment, while bonding prevents destructive voltage differentials from developing between different pieces of gear. The final ingredient necessary for this technique to be effective is a *good station ground.*

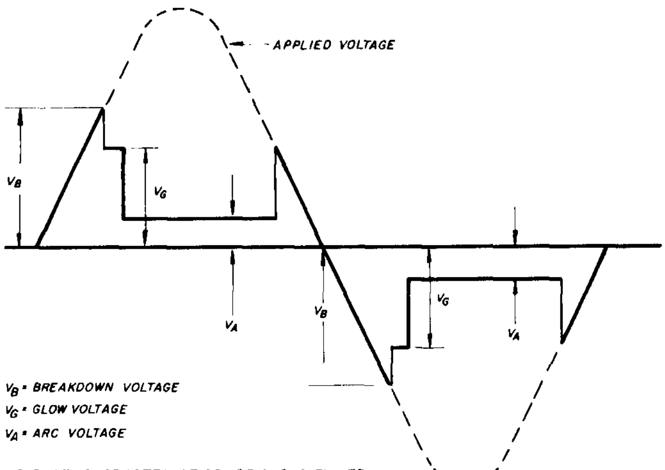
The station ground is second in importance only to the tower ground, and many of the techniques previously covered are applicable. The objective of the station ground is to provide a ground plane at uniform potential for the entire station.

Commercial installations. In commercial installations where an entire building is devoted to the equipment installation, the recommended procedure is to install a buried ground ring around the outside perimeter of the building. This ring is supplemented with ground rods or radials. A second ground bus is run around the inside perimeter of the building. The two ground rings are interconnected at no fewer than four points and at intervals of no more than 15 meters (50 feet). This system is also connected to the power-system ground, and any extensive masses of metal such as water pipes and heating ducts. Finally, the building ground system is connected to the tower ground system. This wire must not be run through a conduit, nor should it be routed near feedlines or other cables from the tower, as it may induce transients into them.

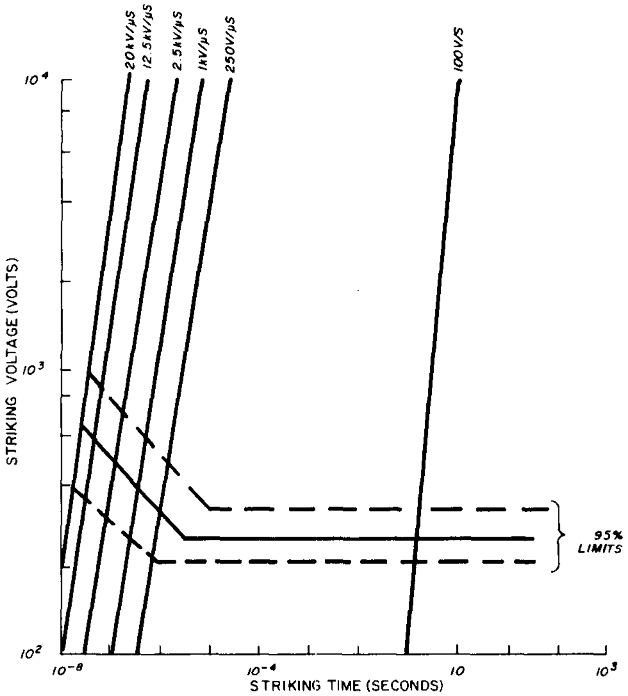
Amateur installations. For a typical amateur station in a home, a commercial-grade station ground would usually be impractical, since all the equipment is usually in one room and near to only a small portion of the building perimeter. The commercial procedures can be modified using common sense to fit the circumstances of any particular installation. The important thing to remember is that a single ground wire to the nearest water pipe is *not* adequate, and every extra bit of grounding adds something to the degree of protection obtained. Whatever modifications are made, the connections between the station ground, tower ground, power system ground, and the water pipes must not be eliminated.

Interconnections and bonding. Once a suitable station ground has been provided, the equipment must be properly connected to it. The shields of all coax feedlines should be connected to the station ground at the point where they enter the building. On each piece of equipment, a ground tiepoint should be selected. This point should be connected to the station ground and also to the ground tiepoint on physically adjacent pieces of equipment. These ground connections should be as short and straight as possible, using 2.6-mm (no. 10 AWG) or heavier copper wire.

It's particularly important on any piece of equipment where a coax feedline terminates from the tower that the point where the feedline connects be used as the ground tiepoint. The requirement for multiple connections of coax and other shielded wires to some pieces of equipment may make it impossible to eliminate secondary ground paths en-



VOLTAGE CHARACTERISTICS OF A GAS FILLED SURGE ARRESTER



maximum ratings	40,000A — 15 pulses of 8/20 waveform at 3-minute intervals 20A RMS ac or dc — for 1 second 10 times at 3-minute intervals
life	10,000 surges typical with 10/1000 waveform at 1000 amps
dc arc voltage	30 vdc, maximum
glow to arc transfer current	0.25 to 1.0 amp
operate delay time	500 nanosecond, maximum
insulation resistance	10^9 ohms
capacitance	
end to end	1 picofarad
end to center	2.5 picofarads
dimensions	8 mm (5/16 in.) dia. × 44 mm (1-3/4 in.) long

fig. 7. Ratings for a TII-16 gas-filled surge arrestor.

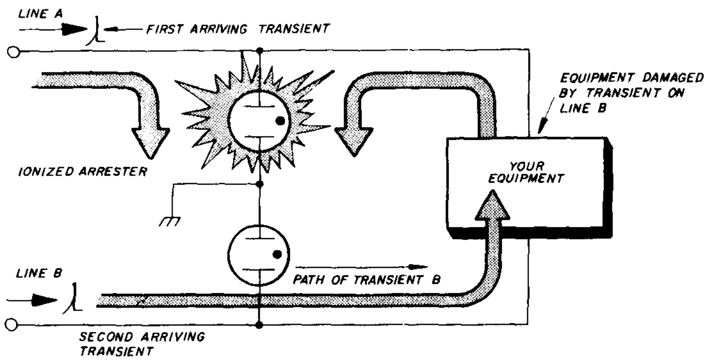
tirely, but careful grounding and bonding will make these secondary paths relatively unattractive to lightning current. The secondary ground path provided by these shielded cables can be made to look like an even higher impedance if the cables are made longer than necessary and the excess length is wound into a coil.

Threewire ac cords. The ground wire in three-wire ac line cords is fine for its intended purpose of preventing shocks resulting from power-line leakages, but it's detrimental as far as lightning protection is concerned. Not only is it neither short nor straight, but its proximity to the other wires in the power cord could couple unwanted surges into the equipment through the power line. When proper grounding and bonding techniques are applied, the ground wire in the three-wire power cord is a secondary ground and no longer needed for its original purpose. Therefore, it should be disconnected for optimum lightning protection.

Dielectric breakdown. The second route to ground at the rig is by dielectric breakdown to power or telephone lines. This is most likely to occur in installations where proper grounding and bonding practices have not been followed. But proper grounding and bonding practices do not completely eliminate the possibility of dielectric breakdown. This is because of the ever-present inductance in any ground wire, together with the fast rise time of the lightning pulse. As already noted, the chassis of all the equipment may rise momentarily to thousands of volts above ground when lightning strikes. The power-transformer primary in each piece of gear is connected to the ac power line, which is at a low impedance and always within a few hundred volts of ground. If the insulation between any of these primary windings and the transformer core, which presumably is chassis mounted, is insufficient to withstand this voltage, dielectric breakdown will occur and the transformer will be destroyed. The same thing can happen with the telephone lines, except that here the phone patch or telephone itself will be destroyed.

Surge suppressors. To prevent this sort of damage a form of bonding must be employed. Obviously the power and telephone lines can't be directly and permanently shorted to ground. Instead, some type of transient voltage surge suppressor must be used. This is a device that's normally an open circuit but which will momentarily break down and provide a low-impedance path across whatever it is connected to in case of a lightning strike. When the surge is over, the device opens the circuit again so that normal equipment operation is unimpaired.

Many types of transient voltage surge supressors are on the market. They offer varying degrees of protection and range in price from a few cents to hundreds of dollars. There are two devices which, when used together, will provide a high degree of protection at relatively low cost. These are the zinc-oxide varistor and the gas-tube surge arrestor. Each has certain advantages and certain limitations; this is why it's recommended that they be used together.[8]



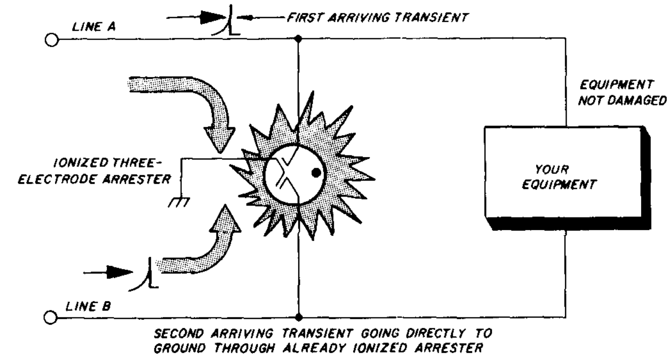


fig. 8. With two-electrode arresters (top), the first arrester to "see" the surge is ionized by that surge and provides an excellent path to ground. The second wire now uses the available ground, but, unfortunately, the path to ground goes right through the equipment to be protected. With three-electrode gas arresters (bottom), if either wire sees a surge, the arrester fires, ionizing the common gas chamber and instantly grounding both wires; there is no damaging potential difference across the equipment.

The zinc-oxide varistor is a device on the market for just a few years, but it has gained wide acceptance in the electronics industry as a transient suppressor. It's available from several manufacturers in a wide range of ratings. The MOV line from General Electric is typical of those available. These devices are designed to protect against transients on the ac power line from lightning and from inductive load switching. Its advantages are response time measured in nanoseconds and response to voltages only slightly above the normal operating voltage of the circuit. Disadvantages are limited maximum-current capability and high "let-through" voltage under high-current conditions (**fig. 6**).

Gas-tube surge arrestors have a firing voltage that depends on the rise time of the transient waveform. Faster rise times result in higher firing voltages. For a lightning waveform having an 8-microsecond rise time, a gas tube can be expected to fire in about one microsecond (**fig. 7**). Once a gas tube has fired, the voltage drop across it is clamped to 30 volts. A gas tube can handle more peak transient current than a varistor, because the low voltage drop results in low power dissipation.

A gas tube will continue to conduct until the applied voltage drops below 30 volts. In an ac-power circuit, this means the tube will conduct for a full half cycle, 8.3 milliseconds at 60 Hz, even though the transient that caused it to fire may have lasted only a few microseconds. This may exceed the long-term power handling capability of the tube, and for this reason gas tubes should always be installed on the *load* side of the fuse or circuit breaker in a power circuit.

Gas-tube surge arrestors are available from TII Corp., Joslyn Electronic Systems, and others. For power-line applications, a three-electrode gas tube is used, with the end electrodes connected to the two sides of the ac supply and the center electrode connected to ground. If the potential between the grounded electrode and either supply electrode becomes sufficient to fire the tube, the entire tube ionizes and all three electrodes are effectively shorted together for the duration of the transient (**fig. 8**). The disadvantage of the gas tube is that it's slower to respond than the zinc-oxide varistor and requires a higher voltage relative to the normal voltage in the circuit before it will fire.

When both a zinc-oxide varistor and a gas-tube surge arrestor are used to protect a piece of equipment, the varistor will take care of short-duration, low-energy transients such as might result from a lightning strike on the power line at a considerable distance from your location. The gas tube will come into play when lightning strikes your tower or the close-by power line. If economic considerations dictate the use of only one of these two devices, the gas tube is the one that should be chosen.

Devices are available using gas tubes that can be inserted between an appliance and the power outlet. This may seem very convenient, but these devices should not be used. The reason is that they are designed to protect against transients coming in on the power line only. We are mainly concerned with protecting against a lightning strike on the antenna or tower. This requires locating the gas tube as close as possible to the power transformer of the equipment to be protected to eliminate the inductance of the line cord from the ground path for the surge. Installation of a gas tube at this location will give maximum protection against a transient either from the power line or antenna. A set of these suppressors should be installed in each piece of equipment connected to the power line for which protection is desired (**fig. 9**).

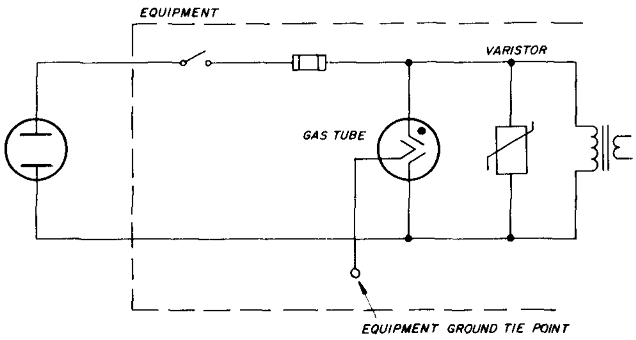


fig. 9. Use of both a gas tube and a varistor for maximum protection.

If your station includes a phone patch, a three-element gas tube should also be installed at the point where the telephone line connects to the patch.

concluding remarks

Lightning protection may appear to be a very complicated subject. But the average amateur station can be well protected without an unreasonable investment of time or money. Remember first and foremost that all of these techniques are nothing more than a means of controlling the path that the lightning bolt will take in its unstoppable search for ground. Work out the required protection system for your individual station with this in mind.

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*A complete bibliography of material on lightning protection is available from *ham radio* and will be sent to interested readers upon receipt of a self-addressed, stamped envelope. **Editor**

ham radio

solar-powered repeater design

Why use antiquated methods to power your machine? Here's a strong case for using solar power — it's inexpensive and effective

In this article we provide the repeater-system designer with useful information about using solar power. The information is based on acceptable criteria for designing commercial solar-power systems and is supported by empirical data obtained in the actual operation of a repeater. The WR5ARO 19-79 solar-powered repeater is on Redondo Peak, New Mexico. It was built using the principles outlined below.

Solar systems to provide power for remote radio sites are not new. Recent advances in technology, and a national concern for conserving our energy resources, have brought recognition to solar power as a useful energy source. For remote sites, solar power is one of the few economically feasible power sources. As with other electronic devices, demand and popularity will reduce the price of solar-powered generators from its presently high level to one within the means of most amateurs. Based on projected technology, cost reductions of 50:1 are possible in the near future. This would put the price of a useful solar electric generator not far above the cost of a good, well-regulated bench-type power supply of equivalent capability. Solar power provides good mechanical reliability (no moving parts), good dependability (with proper array sizing), and an attractive price (considering the alternatives). First, some background.

photovoltaic cells

Silicon solar cells are P-N diodes whose photovoltaic characteristics (ability to produce electricity when exposed to light) have been optimized. Peak-current output occurs when the cell is exposed to direct, unrestricted sunlight that has an intensity of 100 mW per cm.[2]

The rated output current of a solar array is directly proportional to light intensity. Therefore, at 50 mW/cm^2 the array output current is 50 per cent of the peak rated current. The output current capability of a single solar cell is a function of the cell cross-sectional area. Solar-cell output voltage is independent of cell size. Output voltage is constant from 10 mW/cm^2 to 100 mW/cm^2. The equivalent circuit to a solar cell is a constant-current generator with voltage limiting. A typical solar array I/V curve is shown in **fig. 1**.

array sizing

Solar cells are connected in series to provide the required voltage and are connected in parallel to provide the required current. Array specifications are given as a *peak* value. Since the manufacturer has no control over the amount of sunlight available, array specifications are relative to peak sunlight (100 mW/cm^2). It's the responsibility of the system designer to derate the peak specifications to an average value specific for the site.

Most solar arrays come from the manufacturer with enough cells connected in series to be optimum for charging batteries. Consequently, the system designer need be concerned only with the peak current available and the derating necessary to provide some average charging current. The amount of de-

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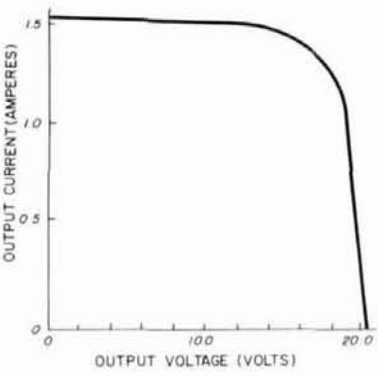


fig. 1. Typical I/V curve of nominal 12V solar array. The curve approximates that of a constant-current power supply with voltage limiting.

rating is a function of the solar-cell insolation at the site.

solar insolation

A number of variables influence the amount of sunlight that strikes the Earth's surface. Among the most important are elevation, cloud cover, atmospheric water content, pollutant level, the sun's incident angle, and the solar-day length. All these factors affect the amount of available sunlight. The amount of sunlight striking the Earth's surface is called insolation. Solar insolation data is given in many different quantities, such as Langleys and kilojoules per square centimeter. The mentioned quantities are usually given for one year; thus the quantities are units of energy. Since solar electric generation is an integrating process, it's permissible to use the average yearly insolation figure to size the solar array. This is true if the average yearly discharge rate of the repeater is also used.

Solar insolation data for Redondo Peak, New Mexico, was found to be 750 kJ/cm^2 per year. This gives an average solar intensity of about 23 mW/cm^2. The average current available at this site is, then, about 23 per cent of the peak current available from the solar array. **Fig. 2** shows average solar intensities for the continental United States.

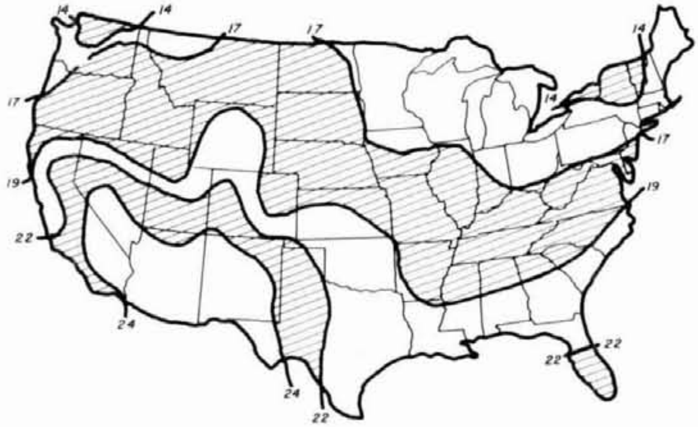


fig. 2. Map of the United States showing average solar intensity in mW/cm^2. Year-to-year variations are less than 10 percent.

storage devices

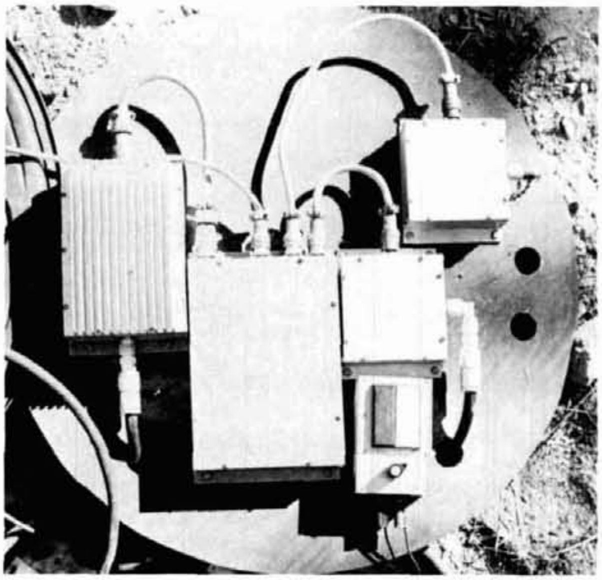
To make power available to the repeater during hours of darkness and foul weather, some type of storage device is used to hold energy collected during hours of daylight and good weather. A battery of some type is almost always used. The capacity of the battery must be large enough to carry the system through extended periods of poor weather and through the shorter days of winter. Battery capacity is relatively independent of array size. Generally, storage capacity is about ten days. The battery is the true power source for the repeater. It should be selected with care.

Co-author WB5RSN makes final angle adjustments to solar array that powers the 19-79 repeater on Redondo Peak, New Mexico (3433 meters, 11,254 feet). The array is mounted at 17 meters (55 feet) to prevent shading from nearby trees.

High-capacity, lead-acid automotive batteries should be avoided. This type of battery is designed to provide large amounts of current for short periods of time. To accomplish this, the battery must have a low internal impedance. High leakage currents occur in this type of design. A good-quality, electronic-grade rechargeable battery would be a better choice. Lead-calcium and gelled-electrolyte batteries, as well as telephone-type "wet" cells, are also good choices. Nickel-cadmiums aren't recommended because of their tempermental characteristics. Remember that most of the vagaries of these cells were discovered in space satellite power systems where solar cells were used to charge them.

voltage regulation

Solar generators normally provide more charging current than a fully charged battery can safely tolerate. To prevent damage to the battery, a voltage regulator is used to limit the charge voltage to a safe level. The circuit shown in **fig. 3** is a simple shunt-type regulator. The series diode prevents the array from discharging the battery during hours of darkness. The diode is also a reverse-bias switch that allows the shunt resistor to absorb the excess power generated. The regulator receives its power directly from the array and therefore does not draw power from the battery. Some battery current is used by the regulator for voltage sampling, but this current is very low.



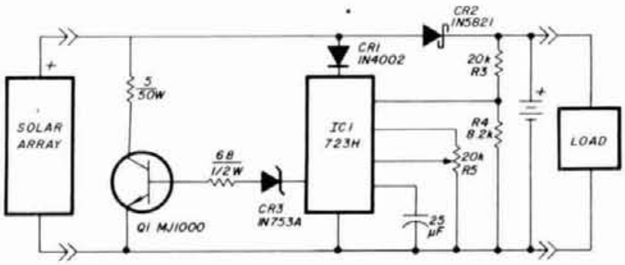
WR5ARO repeater is installed in a sealable 55-gallon barrel. The buried enclosure permits temperature stability. Mountain-top temperatures range from −34C in the winter to 38C in the summer.

Redondo Peak repeater installation. Solar array is mounted near the top of the tower to prevent shading. The repeater electronics are installed underground at the base of the tower.

One of the important characteristics of this type of regulator is its negative temperature characteristic. Simple zener regulators have a positive temperature coefficient, which causes the battery to overcharge in the summer and undercharge in the winter. The opposite characteristics are desirable. Failure modes have been arranged to cause an open circuit in the shunt element, thus permitting the array to charge the battery in the event of a regulator failure. Battery status can be monitored by occasional on-site check-ups or by telemetry.

array orientation

Proper array orientation is required to provide maximum power output during the year. Peak output occurs when the sun's rays are at normal incidence to the array plane. To obtain maximum output, the array is oriented true south (north in the Southern Hemisphere) and inclined from horizontal to an angle approximately equal to the latitude at which the site is located. This angle is then increased a few degrees to optimize the array for the winter months when the days are shorter and the sun is at a lower angle. Solar intensity is constant at all times of the year, but the



NOTES
1. RESISTANCE IN OHMS. K = 1000, 1/2 WATT UNLESS SPECIFIED OTHERWISE.
2. R3, R4, ARE METAL FILM, R5 IS CERMET.
3. CAPACITANCE IN µF.
4. R5 IS SET FOR APPROXIMATELY 14Vdc WITH LOAD AND BATTERY DISCONNECTED.
5. CR2 IS SPECIAL LOW FORWARD DROP DEVICE.

fig. 3. Voltage regulator protects the battery from overcharging. Diode CR2 prevents the battery from discharging through the array during hours of darkness (Courtesy Solar Power Corporation).

solar day is shorter in the winter; hence, the net accumulated energy is lower. Fortunately, the solar array generates more power at lower temperatures, thus offsetting some of the loss.

A tracking array could be designed to follow the sun, but the additional power generated would probably be consumed by the tracking system. Reduced reliability would also be introduced into the system because of the mechanical components. In short, tracking sytems are not a good investment at this time.

equipment selection

The cost of solar systems requires that detailed attention be given to operating power requirements and system power overhead. Ideally, the system would draw no power during standby and would convert all current consumed by the transmitter into rf power. Of course, this isn't possible; therefore, the system designer must minimize repeater standby current and maximize transmitter efficiency. Obviously, vacuum-tube equipment can't be used. Surprisingly, most available solid-state, base-station equipment isn't sufficiently efficient to be considered. Solid-state mobile or portable equipment is a good choice, because it lacks many of the frills found in base-station equipment. Pilot lamps and similar amenities should be powered down or disconnected. Logic circuits should draw a minimum of power. CMOS devices can operate at high-voltage levels with amazingly low current consumption. The WR5ARO identifier is built with CMOS and draws about 50 μA. Likewise, the COR, control circuitry, supervisor, and timers all combined draw less than 1 mA.

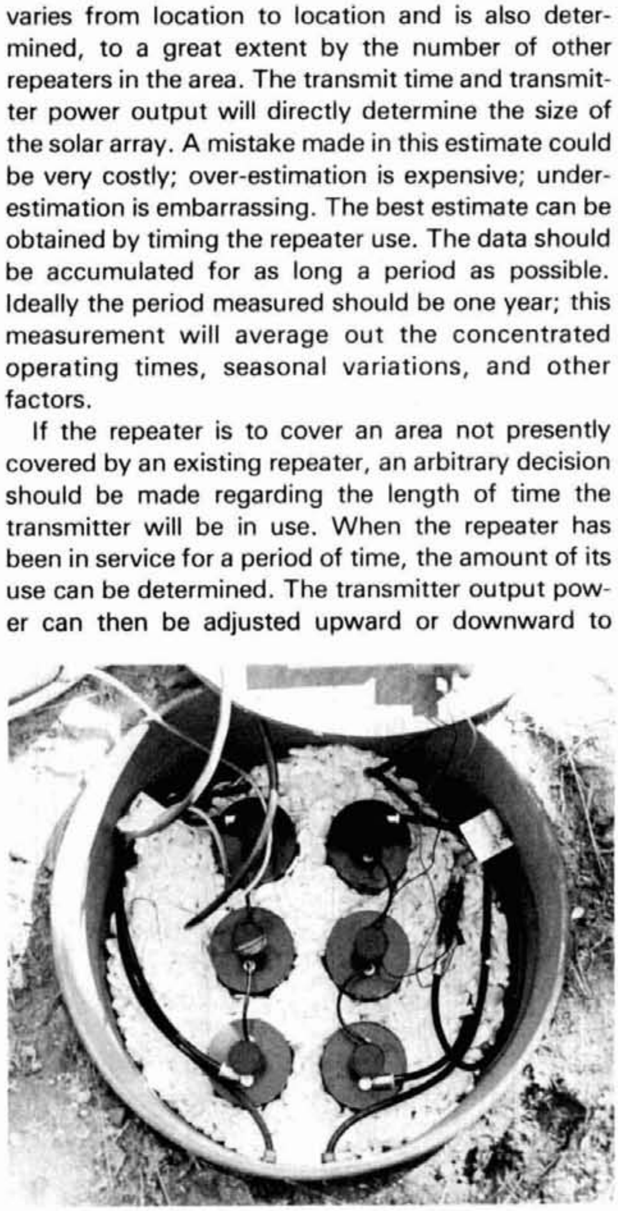
Reasonable numbers to achieve in equipment design or selection are idle currents (total) of 5-20 mA and transmitter efficiencies of 60 per cent. Under these circumstances, the size of the solar array will be a direct function of the transmitter output power.

If operation below 0°C is anticipated, extended temperature devices are required. It's wise to make sure that the circuitry will operate over the expected temperature extremes.

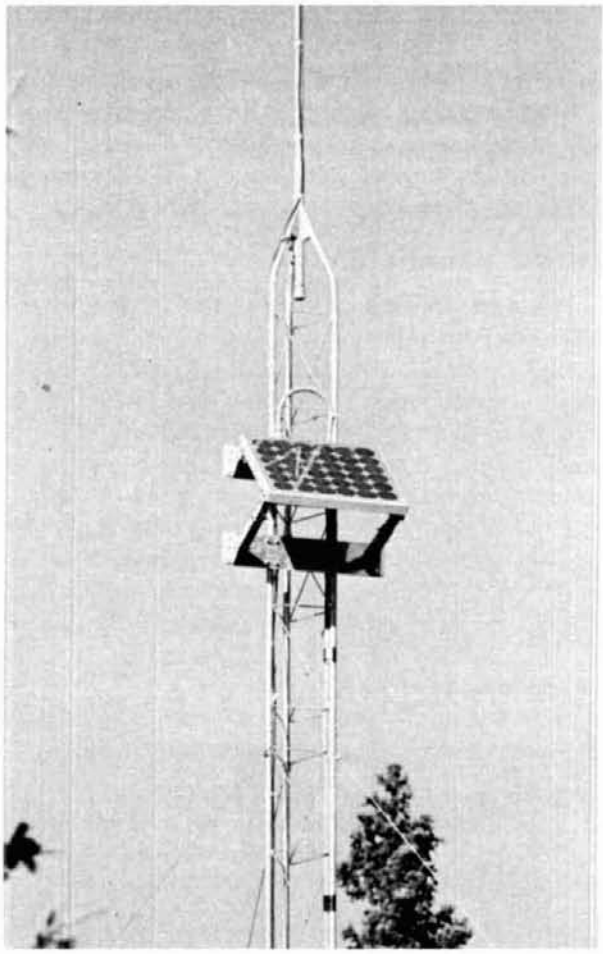
critical parameters

The most difficult data to obtain in designing a solar-powered repeater is the time the repeater is actually on the air. The time that the repeater is used varies from location to location and is also determined, to a great extent by the number of other repeaters in the area. The transmit time and transmitter power output will directly determine the size of the solar array. A mistake made in this estimate could be very costly; over-estimation is expensive; under-estimation is embarrassing. The best estimate can be obtained by timing the repeater use. The data should be accumulated for as long a period as possible. Ideally the period measured should be one year; this measurement will average out the concentrated operating times, seasonal variations, and other factors.

If the repeater is to cover an area not presently covered by an existing repeater, an arbitrary decision should be made regarding the length of time the transmitter will be in use. When the repeater has been in service for a period of time, the amount of its use can be determined. The transmitter output power can then be adjusted upward or downward to



Duplexer cavities are mounted in a barrel. Styrofoam packing chips slow rapid thermal changes. Cavities are copper-plated steel to minimize any detuning due to thermal stresses.



Solar panel is oriented true south and is inclined about 40 degrees for horizontal. Latitude of the repeater is approximately 36 degrees north. An additional four degrees of inclination optimizes the array output during winter when the sun is lower in the sky and days are shorter.

match the time use to the solar array. WR5ARO was designed to handle one-half the local traffic load that's presently divided between four local, wide-coverage repeaters. At its present power output, the repeater can comfortably provide 35 hours of operation per week.

Battery capacity is an important consideration in system design. The capacity figure will determine how long the repeater will operate when the solar array is not charging, or when the repeater is being used at a current rate greater than can be supplied by the array. It's necessary for the solar array to supply more current than the repeater will consume. It's not necessary for the array to fully charge the battery each day. The system can tolerate some deficit as long as the battery is not damaged by freezing or any peculiarities typically inherent to the type of battery used.

design procedure

Designing a solar power supply for a repeater isn't difficult. Remember that solar energy collection is a cumulative process. Its occurence is very regular and very predictable. Year-to-year variations are less than 10 per cent. Repeater use must be averaged to fit the collection criteria. The battery capacity is selected to be adequate to equalize the short-term variations in repeater use and local weather phenomena, which are highly unpredictable. The steps necessary for design are as follows:

1. Determine solar insolation for the proposed site. Source data can be obtained from sources listed at the end of this article. **Fig. 2** may be consulted directly.

2. Determine continuous idle current, multiply by 24 hours to determine the daily idle current ampere-hours.

3. Determine the transmitter current; multiply it by the transmitter on time to determine the daily average transmitter ampere hours.

4. Average the repeater load over a 24-hour period. Then, divide the idle ampere-hours plus the transmit ampere-hours by 24. This is the average load current. This number must be less than the average charge current as supplied by the array.

5. Determine the peak-panel output. The average solar intensity should be found (**step 1**) and divided into the average daily load current. Remember that if the average intensity is 22 mW/cm^2, then the average current available from a solar panel will be 22 per cent of the peak.

6. Calculate the "no-sun" storage requirement of the battery. Ten days of storage is an average number. Multiply the total ampere-hour load (**steps 2** and **3**) by ten to obtain the battery capacity.

Note that nothing has been said about battery-charging efficiency. Battery efficiency cancels, because the charging voltage is greater than the discharge voltage. The solar array provides the additional charging voltage required by design, with no sacrifice in performance. This assumption is valid if the internal leakage of the battery is not great (less than 3-5 per cent per month).

conclusions

Solar power is useful in providing adequate power to operate a radio repeater if care is used in designing the system. The designer has a wide latitude of options available. Enough considerations have been given to demonstrate that gross overdesign of a solar-power generator is not necessary. Attention to details and careful consideration of all available options will produce an economical design.

The Redondo Peak repeater has been in full solar-power operation since June 18, 1977. There has been no down time. A system checkout on December 16, 1977, showed that the solar array was generating its rated power output and the battery was fully charged.

The cost of the solar generator, when averaged over its 20-year life, comes to about $35 per year. This number compares quite favorably with the price charged to many mountain-top customers for similar power. As the price of solar power drops, so will the yearly cost for power generated by this means. In today's energy-cost spiral, solar power will become very attractive in the near future.

bibliography

1. "Solar Electric Generator Systems, Principles of Operation and Design Concepts," Solar Power Corporation, 5 Executive Park Drive, North Billerica, Massachusetts 01862.
2. "The National Atlas," U. S. Government Printing Office.
3. "Annual Solar Radiation in kJ/cm^2," map from Sensor Technology, Inc.

appendix

WR5ARO Specifications

idle current (mA)	12.
transmit current (A)	1.07.
operation	available 24 hours (open repeater).
design average daily use	2.5 hours per day.
transmitter power output (W)	9.5.
effective radiated power (W)	35.
solar array source	Solar Power Corporation E-01-369-1.5 (1.5 A peak output).
battery source	Globe Union Gel Cel 40 A-h (2 each GC12200).
environmental characteristics	elevation 3.43 km (11,254 feet). temperature −34°C to 38°C (−30F to 100°F). solar insolation 750 kJ/cm^2/year. average solar intensity 23.7 mW/cm^2. rainfall 46 cm (18 inches) per year. snowfall 91 cm (36 inches) per year.

Sample Calculations Using WR5ARO Data

Step 1 Solar insolation data = 750 kJ/cm^2/yr = 23.7 mW/cm^2.

Step 2 Continuous load = 0.012 A × 24 hrs = 0.288 A-h.

Step 3 Intermittent load = 1.07 A × 2.5 hrs = 2.675 A-h.

Step 4 Daily average load = 2.963 A-h/24 hrs = 0.123 A.

Step 5 Peak panel output = 0.123 A + 10 per cent/(23.7 mW/cm^2/100) = 0.570 A peak.

Step 6 Storage Capacity = 2.963 A-h × 10 days = 29.63 A-h battery. (Add some additional capacity to prevent freezing in the winter).

ham radio

universal digital readout

A universal digital readout system featuring reduced ambiguity, high input frequency, and low power consumption

This article describes a relatively simple design for a digital dial that evolved over several years of building and improving. For easy home duplication, it uses a minimum number of specially selected components. It is adaptable to virtually all types of shortwave equipment. Options are also described for low-power operation and reduced last-digit flicker.

The counter is connected to the vfo of the equipment and preset to the i-f, or the complement of the i-f. It is wired to count up or down depending on the internal frequency scheme of the equipment. It is even possible, without knowing these parameters, to set up the counter using only one calibration point and check whether the frequency indication moves in the right direction.

This counter has a 35-MHz capability, and thus covers the entire conventional shortwave range (3 to 30 MHz). It is therefore possible to measure the frequency that is generated by the premixing scheme in Drake equipment.

counter components

counter and readout. The basic four-digit counter consists of four low-power Schottky BCD counters, the 74LS190 (see **fig. 1**). The BCD outputs of the counters are connected to special LED readouts which contain an internal latch/decoder. The readouts, HP-type 7300, are somewhat expensive. However, for the home brewer they immensely simplify construction.

As mentioned, the counter reads only the four most significant digits, since the MHz digits have always been read from the band switch in the past. Plus, the complexity of added digits might make the job more than the average ham would want for a home project. The motto here is keep it simple.

time base. The time base is a very simple circuit. It consists of a single IC (CD4060) and a crystal, a trimmer capacitor, and a resistor. The IC contains the necessary amplifiers for a crystal oscillator and 14 divide-by-2 stages. At the output of the last stage, labeled Q_{14}, the oscillator frequency has been divided by a factor of 2^{14}. Starting with a 409.6-kHz crystal, the final output is a 25-Hz squarewave. This output, plus the 50-Hz squarewave from Q_{13}, are used to generate the necessary counter timing pulses.

counter timing. Operation of a counter generally requires various timing pulses to control the counter. In a conventional counter, the count gate provides a precisely timed interval which allows the number of counts admitted to be equal to the frequency of the signal. Since frequency is measured in terms of events per second, this gate is always a fraction of a second. Or, in this case, where we want to read to hundreds of Hz, the gate is exactly 0.01 second long. Other pulses are required for presetting the counter to a fixed starting number (frequency) and for transferring the final count to the readouts.

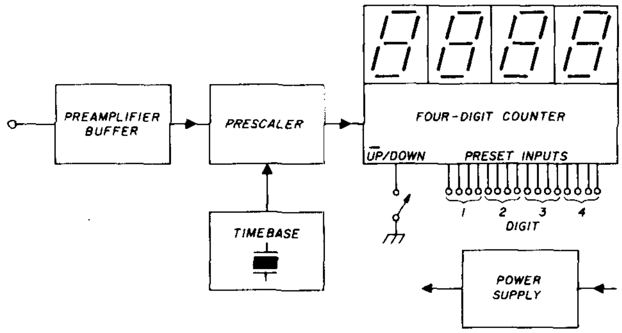


fig. 1. Simplified block diagram of the universal digital readout.

The count gate, the display, and the preset pulse are all derived from the two squarewaves provided by the time base (see **fig. 2**). The first two gates connected to the CD4060 buffer the CMOS outputs of the time base.

preamplifier. A preamplifier, though not always necessary, is a good idea. It not only increases sensi-

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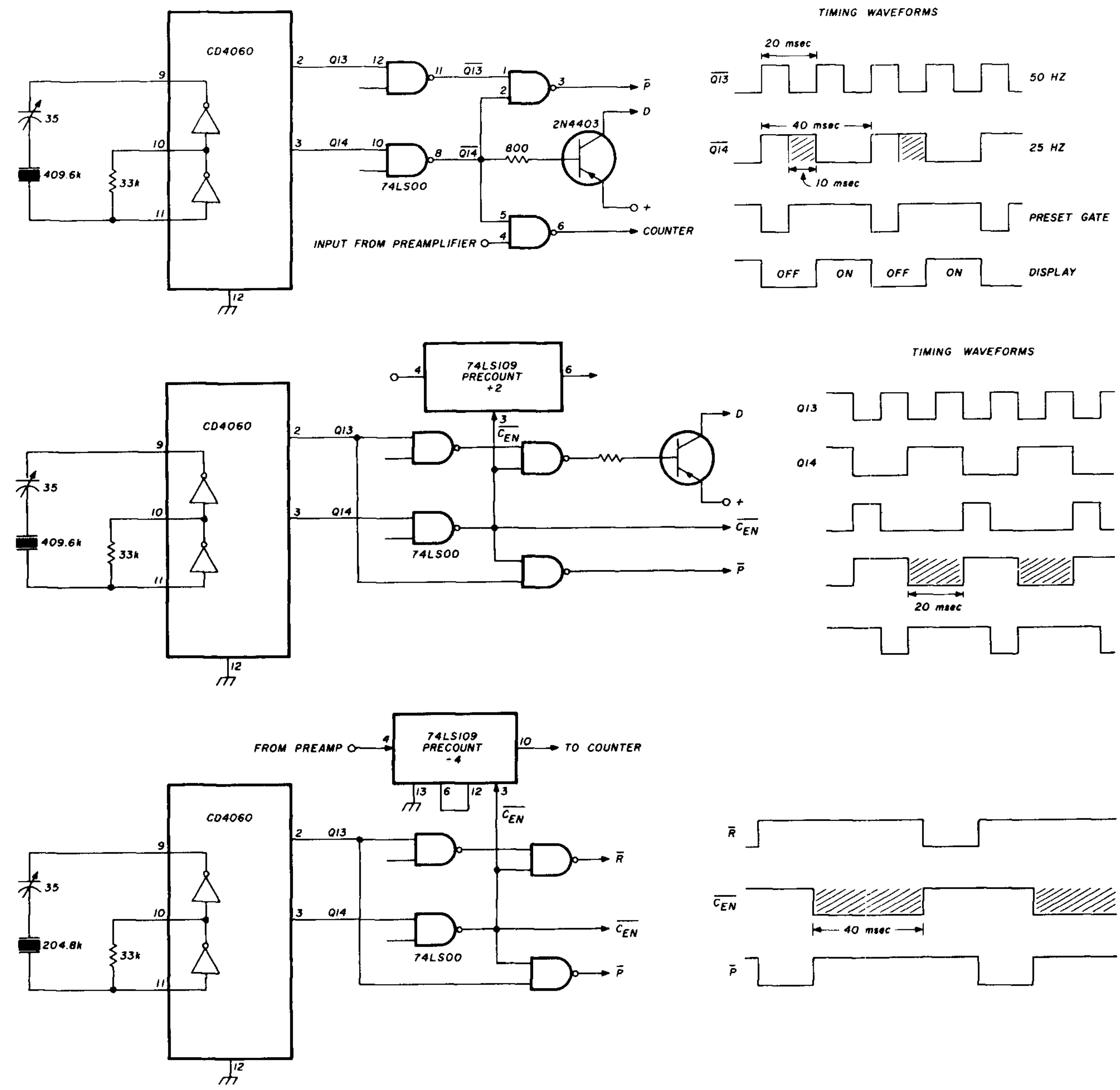

fig. 2. Schematic diagram of the three different timebase versions. In A, the count gate is actually enabled for 20 mS, but since the counters are held in the preset state for the first 10 mS, the time that the counters are allowed to be clocked is 10 mS. The low-power version, B, runs the display at a 25 per cent duty cycle. For the low-ambiguity version, C, the crystal is changed to 204.8 kHz, effectively quadrupling the count gate to 40 mS.

tivity but also acts as a buffer, reducing possible spurious responses in the receiver generated by the timing pulses. A single transistor, as shown in **fig. 3**, is used in common-emitter configuration. The sensitivity is better than 50 mV RMS from 100 kHz to 30 MHz. The maximum voltage is about 1 volt RMS.

power supply. The digital dial, using mainly TTL-devices, requires a 5-volt dc power source. The current, depending on the desired version, will range from 170 mA for battery-powered equipment to 500 mA for the low-ambiguity 100 per cent display version.

last digit ambiguity

The problem last-digit ambiguity arises from the fact that the count gate, as generated from the crystal oscillator, is not synchronized to the incoming frequency. The gate will sometimes accept an additional count, changing the digit, for example, from a 5 to a 6. This is the well-publicized ±1 digit ambiguity that digital counters exhibit. One way of overcoming this ambiguity is to increase the number of digits counted, yet only display a limited number. This, in effect, is the same as simply covering the blinking digit.

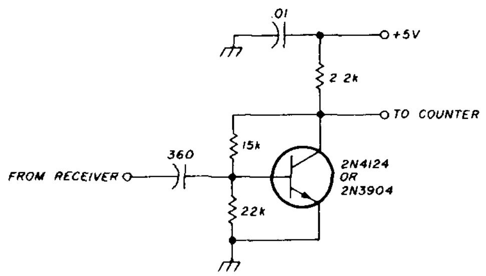

fig. 3. Schematic diagram of a simple buffer/preamplifier which can be used between the counter and the receiver. Sensitivity is better than 50 mV RMS.

The conventional counter can achieve the additional counts only by lengthening the gate; if we consider another decade, the count gate would be 0.1 second, reducing the counter's final update rate to only 6-8 Hz. It is still possible, using the memory capability of the readout, to obtain a blink free and *almost* instantaneous update when turning the dial of the VFO.

However, it is not actually necessary to add a complete decade to the counter; any integer will do. How then is the ambiguity affected by the addition of the new counter? *In reality, it never goes away. What* does happen is that the probability that the ± 1 ambiguity will occur is reduced by the reciprocal of the additional factor. For example, if you add a divide-by-two, the probability will be reduced by 1/2, or 50 per cent; for a divide-by-four it will be 1/4, or 25 per cent. However, for this reduced probability, there is a price that must be paid. The count gate will have to be *lengthened* by the same factor.

programming

To program the counters, the individual load lines, labeled D_A, D_B, D_C, D_D in **fig. 4**, are connected according to the required BCD code. To program a 5 into a particular counter, ground the data lines for D_B and D_D. The other data lines may remain open or connected to +5 volts.

A *simple scheme using a single-pole, double-*throw switch, as shown in **fig. 5**, can be used to preset the counter to two different starting frequencies. For more than two positions, a multiple-deck switch would be required.

counter options

standard version. This is the simplest form of the digital dial, with no precounting to reduce ambiguity. The time base generates a 0.01-second gate, giving a readout to the nearest 100 Hz. The display is updated at the rate of 25 Hz, with a display duty cycle of 50 per cent. Power requirements are 5 volts at 300 mA.

low-power version. In this version the time base output is slowed to 0.02 second, permitting the addition of a single divide-by-two counter which reduces the last digit flicker to 50 per cent. The display duty cycle is also reduced, to 25 per cent, giving a somewhat dimmer, but still quite visible, display. The update rate is still 25 times per second. Power consumption is under 1 watt (170 mA at 5 volts).

low-ambiguity version. For a little added complexity, this version is the most useful for fixed-station use. The time base is further slowed to 0.04 second by substituting a 204.8-kHz crystal for the 409.6 crystal. With this change, a divide-by-four prescaler, which reduces the last digit ambiguity to 25 per cent, can be used.

Since the update rate is now only 12.5 Hz, an intolerable flicker would occur if the display were switched at that rate. To eliminate the flicker, the latch in the

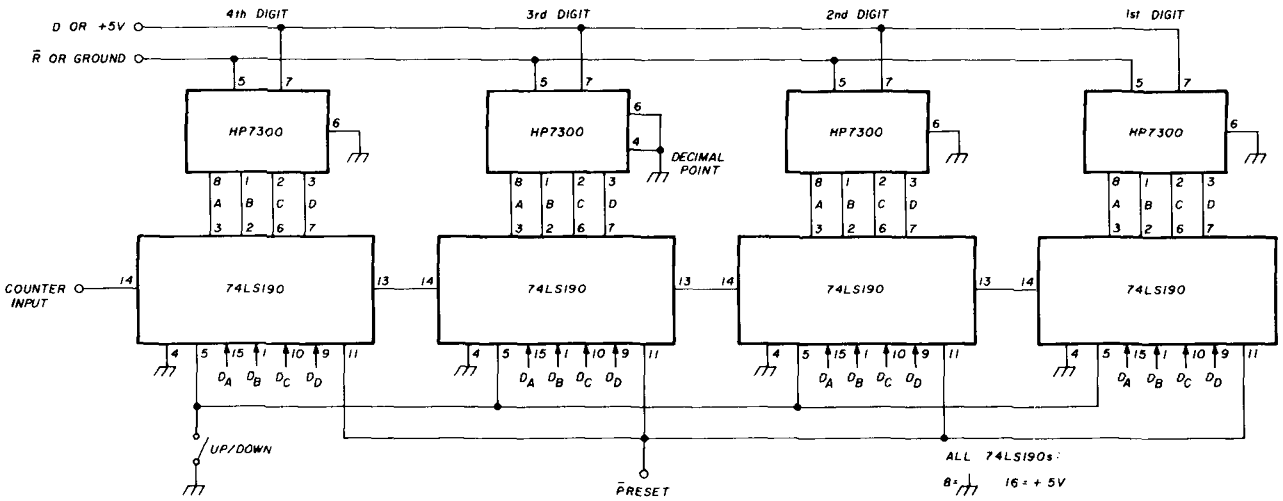

fig. 4. Schematic diagram of the four 74LS190 counters and the HP-7300 LEDs. The R̄ line, connected to pin 5 of the 7300s, is used to strobe the latches when used in the low-ambiguity version. Pin 5 of the counters is taken low for up counting, and can be left open to count down.

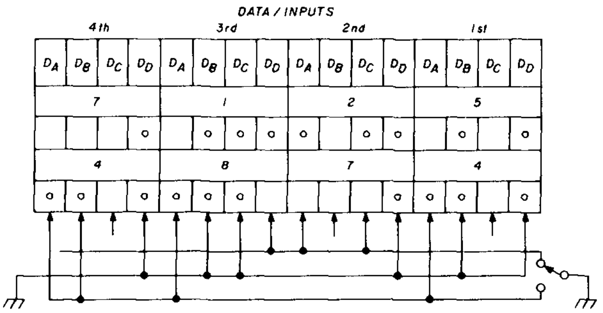

fig. 5. Example of using a single-pole double-throw switch to select different preset programming. Those lines that are always high are left open, while those that are always low are held low. The switch changes the level into the inputs depending upon the preset value.

7300s is also strobed, giving a 100 per cent display cycle. Even with the bright display power consumption is quite reasonable, 500 mA at 5 volts.

checkout and calibration

With an adequate power supply connected to the counter, checkout and calibration can be completed in a few simple steps:

1. Program the preset inputs according to the required BCD input. For a quick check, the number 7 can be programmed into the counter by grounding all pin 9s.

2. Apply a stable rf signal to the input of the preamplifier. The counter should count either up or down, depending upon the input to the control pin.

3. Next, apply a signal of known frequency (crystal calibrator, for instance). Check the displayed frequency against the input, the preset, and whether the counter was programmed to count up or down.

parts list

4 ea 74LS190
4 ea *HP 7300 Numerical Indicators*
1 ea CD 4060A
4 ea Resistors 2.2k, 15k, 22k, 33k, 1/4 W
5 ea Ceramic Cap 2x.1, 800p, 360p, 50p 50V
1 ea Trimmer Cap 35p
1 ea Rectifier Diode 1A, 50V PIV
1 ea LM 309 K IC
1 ea Electrolytic Cap 250 μF 15V
div. HW, sockets, chassis, etc.

Plus:

Standard Version
409.6 kHz crystal
PNP trans.
820Ω Res.

Low-Current Version
409.6 kHz crystal
PNP trans.
74LS109 IC
820Ω Res.

Low-Ambiguity Version
204.8 kHz crystal
74LS109 IC

If all readings are correct, the counter can then be permanently connected to the receiver.

bibliography

Gerd, Schrick, WB8IFM, "Digital Readout Variable-Frequency Oscillator," *ham radio*, January 1973, page 14.

ham radio

the Oscar Calcu-puter

If you are adventuresome enough to attempt automatic antenna tracking of Oscar, but don't have the necessary bucks to tie into a computer, this article should be just what you need. Even if mathematics isn't your strong suit, don't get discouraged, read on. The formulas for tracking Oscar are not that tough, especially if you apply one of the inexpensive hand calculators. Unfortunately, the main disadvantage of the hand-calculator method is the need to constantly manipulate the buttons, even for information for the Oscar pass. This article will explain my method of solving this problem — automating a small hand-held calculator. I'll even explain a few ideas for making a complete steering system. That way all you have to do is enter the equator crossing longitude, punch a button when Oscar crosses the equator, and from there on it's automated all the way.

program explanation

The terms I've used in the program (see **fig. 1**) have been summarized in **table 1**. In addition, I've assigned line numbers to each program step to make it easier to follow. Actually, the program is divided into six separate parts, each solving one of the following equations:[1]

$$Lat(T) = sin^{-1}[0.9790 \cdot sin\,(3.1319T)] \tag{1}$$

$$Long(T) = cos^{-1}\{cos(3.1319T)/cos[Lat(T)]\} + 0.25T + Lo \tag{2}$$

$$D = cos^{-1}\,(sinA sinB + cosA cosB cosL) \tag{3}$$

$$Az = cos^{-1}[(sinB - sinA cosD)/cosA sinD] \tag{4}$$

$$El = 90 - tan^{-1}\,[4867 sinD/(4867 cosD - 3957)] \tag{5}$$

$$M = (4867 cosD - 3957)/cos(90 - El) \tag{6}$$

Steps 004 through 030 solve **eq. 1**, 031 to 058 **eq. 2**, 059 to 108 **eq. 3**, 109 to 148 **eq. 4**, 149 to 191 **eq. 5**, and steps 192 to 203 for **eq. 6**. Each step is actually a single key-stroke on the calculator. There are several steps that should be briefly explained. This might eliminate program questions as you follow the equations through the program.

Step 000 represents the unit being turned on. In addition, other circuitry resets the external logic back to a common starting point. The one-shot multivibrator which performs the reset function also enables the clock gate. Anytime the clock gate is enabled, the sequencer is allowed to advance to the next program step. If the gate is disabled, the program will stop on that particular step. This is an important feature, as I'll explain later on.

Steps 001 through 003 merely clear the calculator of any previous computations or stored answers. The next two steps shift the calculator into four decimal place readout. The calculations are actually done to the limit of the ICs involved, and the answers rounded. For all program steps you'll notice a listing for type of entry. This notation is explained in **table 2**.

Program steps 013 and 014 cause the time since the satellite crossed the equator to be entered into the calculator. This is entered as even minutes and results in a readout for antenna azimuth, elevation, and distance to the satellite for each minute of the pass. In theory, at least, if you provide the correct initial data and accurately enter the time, the calculator could provide information for tracking for the next pass, or even several later passes. This is limited only by the accuracy of the entries you make, and could easily be updated.

The last unusual steps are 038 and 039. In some calculations it is easier to find a denominator before

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fig. 1. Program solved by the Calcu-puter. When broken into parts, this program will solve the six equations necessary to track Oscar.

Type of Entry	Program Step	Key
*	000	
K	001	DPS
M	002	CLR
M	003	CLX
D	004	DPS
D	005	4
Co	006	3
C	007	DP
Co	008	1
Co	009	3
Co	010	1
Co	011	9
M	012	ENT
Ti	013	T_{tens}
Ti	014	T_{ones}
Me	015	STO
Me	016	1
M	017	X
Me	018	STO
Me	019	2
T	020	SIN
C	021	DP
Co	022	9
Co	023	7
Co	024	9
Co	025	0
M	026	X
T	027	INV
T	028	SIN
Me	029	STO
Me	030	3
T	031	COS
Me	032	STO
Me	033	4
Me	034	RCL
Me	035	2
T	036	COS
M	037	÷
K	038	DPS
K	039	1/X
T	040	INV
T	041	COS
C	042	DP
Co	043	2
Co	044	5
Me	045	RCL
Me	046	1
M	047	X
M	048	+
S	049	EQX 100
S	050	EQX 10

Type of Entry	Program Step	Key
S	051	EQX 1
C	052	DP
S	053	EQX .1
S	054	EQX .01
M	055	+
Me	056	STO
Me	057	5
M	058	CLX
Cq	059	LAT 10
Cq	060	LAT 1
C	061	DP
Cq	062	LAT .1
Cq	063	LAT .01
Me	064	STO
Me	065	6
T	066	SIN
Me	067	STO
Me	068	7
Me	069	RCL
Me	070	3
T	071	SIN
Me	072	STO
Me	073	8
M	074	X
Me	075	STO
Me	076	9
M	077	CLX
Me	078	RCL
Me	079	6
T	080	COS
Me	081	STO
Me	082	1
M	083	ENT
Me	084	RCL
Me	085	4
M	086	ENT
Cq	087	LONG 100
Cq	088	LONG 10
Cq	089	LONG 1
C	090	DP
Cq	091	LONG .1
Cq	092	LONG .01
M	093	ENT
Me	094	RCL
Me	095	5
M	096	-
T	097	COS
M	098	X
M	099	X
Me	100	RCL
Me	101	9

Type of Entry	Program Step	Key
M	102	+
Me	103	STO
Me	104	2
T	105	INV
T	106	COS
Me	107	STO
Me	108	3
Me	109	RCL
Me	110	7
M	111	ENT
Me	112	RCL
Me	113	2
M	114	X
Me	115	STO
Me	116	5
Me	117	RCL
Me	118	8
M	119	ENT
Me	120	RCL
Me	121	5
M	122	-
Me	123	STO
Me	124	5
M	125	CLX
Me	126	RCL
Me	127	1
M	128	ENT
Me	129	RCL
Me	130	3
T	131	SIN
Me	132	STO
Me	133	6
M	134	X
Me	135	RCL
Me	136	5
M	137	÷
K	138	DPS
K	139	1/X
T	140	INV
T	141	COS
D	142	DPS
D	143	0
R	144	GoSUB1
D	145	DPS
D	146	4
Me	147	STO
Me	148	4
M	149	CLX
Co	150	4
Co	151	8
Co	152	6
Co	153	7

Type of Entry	Program Step	Key
Me	154	STO
Me	155	7
Me	156	RCL
Me	157	2
M	158	X
Co	159	3
Co	160	9
Co	161	5
Co	162	7
M	163	-
Me	164	STO
Me	165	8
Me	166	RCL
Me	167	7
M	168	ENT
Me	169	RCL
Me	170	6
M	171	X
M	172	÷
T	173	INV
T	174	TAN
Me	175	STO
Me	176	9
Co	177	9
Co	178	0
M	179	ENT
Me	180	RCL
Me	181	9
M	182	-
Me	183	STO
Me	184	1
D	185	DPS
D	186	0
R	187	GoSUB2
D	188	DPS
D	189	4
Me	190	STO
Me	191	6
Me	192	RCL
Me	193	8
M	194	ENT
Me	195	RCL
Me	196	1
T	197	COS
M	198	÷
D	199	DPS
D	200	0
R	201	GoSUB3
Me	202	STO
Me	203	5
**	204	STOP
**	205	END

its associated numerator. This leads to dividing the denominator (d) by the numerator (n), or d/n. To get the correct answer (n/d vs d/n), the reciprocal key ($1/X$) is used after the division answer has been obtained. On my calculator, I have to use the DPS key to access the $1/X$ function. This accounts for the use of two steps.

circuit description

The heart of the automating device is a 1702A EPROM (see **fig. 2**). The program, from **fig. 1**, is entered into the PROM such that the outputs, when decoded, will electrically press the appropriate keys on the calculator. For general use, a RAM would be more appropriate, but I wanted to solve one specific problem: the equations necessary to track Oscar.

The 1702A PROM has 256 distinct address locations, enough to handle the 205-step program. Each step of the program represents a sequential binary address in the PROM. As seen in the schematic diagram, 74193, 4-bit binary counters are used to sequentially address the 1702A. I decided to use the

table 1. Definition of terms used in the Calcu-puter program.

(T)	Time in minutes since EQX
Lat(T)	Satellite sub-point Latitude in degrees at (T)
EQX	Satellite equator crossing reference
Long(T)	Satellite sub-point Longitude in degrees at (T)
Lo	Satellite sub-point Longitude in degrees at EQX
D	Great Circle distance Stn to Sat in degrees
Stn	Station location (your QTH)
Sat	Satellite location
A	Latitude of Stn in degrees
B	Lat(T)
Ls	Longitude of Stn (QTH) in degrees
L	Ls - Long(T)
Az	Azimuth bearing for antennas (from true North)
E or El	Elevation bearing for antennas (from horizon upward)
M	Distance in statute miles from Stn to Sat (true position)

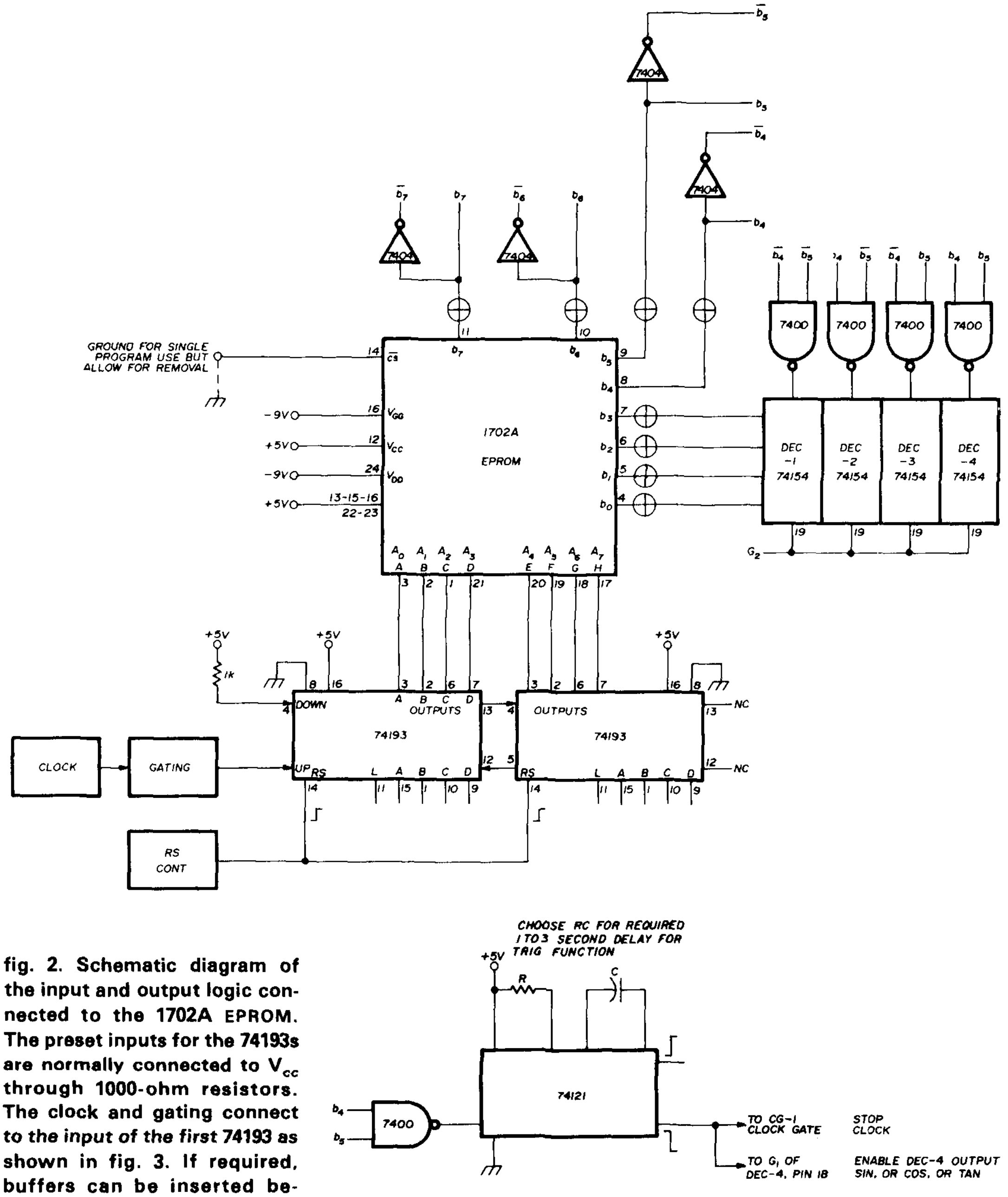

fig. 2. Schematic diagram of the input and output logic connected to the 1702A EPROM. The preset inputs for the 74193s are normally connected to V_{cc} through 1000-ohm resistors. The clock and gating connect to the input of the first 74193 as shown in fig. 3. If required, buffers can be inserted between the outputs of the PROM and any external devices. The actual pin numbers on the inverters and NAND gates have not been shown to allow flexibility in other systems. To incorporate the trigonometric delay feature, a 74121 is inserted between the decoder and the 7430 NAND gate.

74193 instead of the 7493, taking advantage of the preset capability. This means that the program can be started at any spot by simply entering the correct starting address into the data inputs and momentarily taking load line low. If this capability is not desired, the 7493 could be used.

Eight output lines are available on the 1702A. For direct calculator control, I've only used six of the available outputs. The first four outputs, b_0 to b_3, are used as normal addresses for 74154 one-of-sixteen decoders. The outputs are simultaneously applied to all four decoders. The b_4 and b_5 outputs from the 1702A are also decoded and used to select the appropriate 75154.

The final two outputs, b_6 and b_7, are used as a program stop and program halt. When step 204 is addressed, the output from the PROM will be 01000000. The high level from the b_6 output is detected and used to stop the program until a new "minutes" time is entered. This is one of the different means of disabling the clock gate. Output b_7 is programmed in a like manner to provide a high output at step 205. This output will stop the program, regardless of the minutes timer.

equator, the first flip-flop is set, which in turn sets the second flip-flop. Having both set will enable the clock gate.

I have tried to divide the decoders into a logical order, with DEC1 using the binary codes for number 0 through 9, to directly decode the number information.

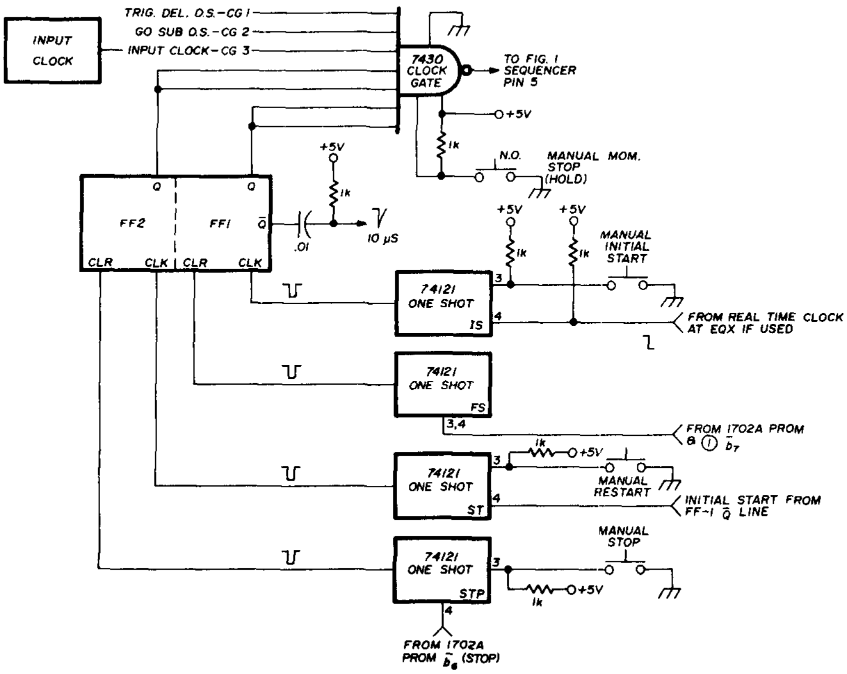

fig. 3. Schematic diagram of the clock gating logic. The actual input clock frequency will depend upon the speed at which your calculator can do the computations. For initial testing, it could be as slow as 1 pulse per second.

Two flip-flops are used for clock gating. As seen in **fig. 3**, the one-shot multivibrators receive the various start and stop commands. The pulses are then used to trigger the flip-flop into the desired states. In addition, provisions have been made to interface a real-time clock to signify the equator crossing. When you initiate the start command as Oscar crosses the

table 2. Type of entry notation used in the program.

*	Power on, reset timer (T) and program sequencer to 000, and all readouts to zero
M	Machine function — CLX, ENT, +, − etc.
D	Change of decimal point location
Co	Constants defined by Oscar and put into PROM
Cq	Constants defined by your QTH and put into only your PROM
Ti	Entry from timer output
Me	Storage or Recall function to/from memory and number
T	Trigonometric function (and added delay trigger)
K	Keyboard shift function (SHIFT that is not DPS)
S	EQX Longitude entry by switches
C	Constant (decimal point entry)
R	Readout Sub-routine function (external to calculator)
**	Operational system command (Stop, End)
STOP	Halts calculator until next timer period enters
END	Detects maximum period of pass elapsed — full stop

tion. The keys for the four basic math functions are also included in DEC1. As seen in **fig. 4**, outputs from the decoders are used to drive open-collector buffers. The buffers then drive the reed relays which are connected across the calculator keys. It is imperative that the relays have a very high resistance across the open contacts, and also a very low closed-contact resistance.

A complete listing of the respective decoder addresses is given in **table 3**. Note that the first address in DEC1 does not have an associated function. This is to prevent a problem when the step 204 and step 205 commands are initiated. If the address were used, you would have a simultaneous key closure in addition to either a stop or halt command.

You'll notice that the only functions in DEC4 are the trig functions. This was done for a very specific reason. In most calculators, depending upon the IC set used, a trigonometric operation will take longer to perform than a basic math function. This was also true in the calculator I used. To overcome the problem, I needed some method of momentarily stopping the program until the calculator had completed the trig operation. Otherwise, the program might have advanced several steps without correct data from the trig operation. By placing these operations in DEC4, detecting any 0011xxxx number will automatically indicate that a trig function is present. After the function has been detected, a one-shot disables the clock gate long enough for the calculator to do the computation.

Instead of using this program-delay technique, the time between steps could be made long enough to allow for a trigonometric operation, but this will considerably slow down the time necessary to perform all the calculations. Using the one-shot requires a few more parts, but the tradeoff is worthwhile, considering the time saved.

The GoSub routines, listed in DEC3, are used to output the data from the calculator to external readouts and other external processing. Gating for the GoSub routines is shown in **fig. 5**, with a quasi-schematic diagram of the readout system shown in **fig. 6**.

On the subject of the GoSub routines, you'll notice a DPS 0 step just before each GoSub step in the program. This truncates the display to eliminate any numbers to the right of the decimal point, and also shifts the answer to position the units digit on the extreme right of the display.

In the multiplexed displays (as used in my calculator), the same segments of each display are tied together, with a digit strobe activating the appropriate digit. The Calcu-puter, as I've aptly named it, is interfaced to external readouts by connecting the segment information lines and the data strobe lines to external latches. **Fig. 6** is not an absolute schematic diagram since the voltage levels and required interface, will differ between calculators.[2] You'll also notice the use of digital information to indicate the actual antenna position. This information, combined with the Calcu-puter information, nicely lends itself to completely automated antenna control.[3]

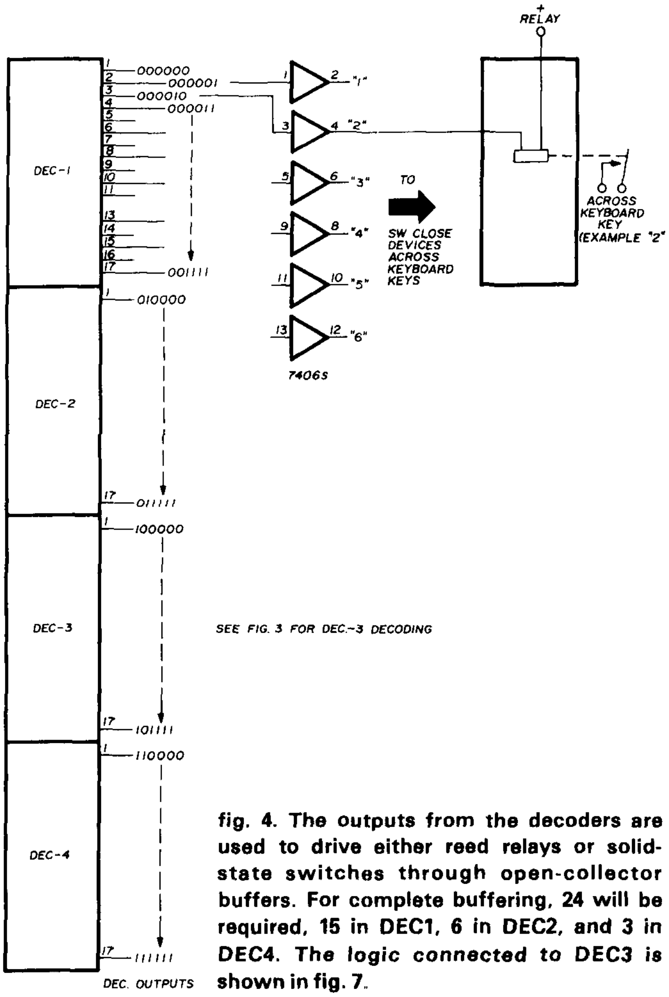

fig. 4. The outputs from the decoders are used to drive either reed relays or solid-state switches through open-collector buffers. For complete buffering, 24 will be required, 15 in DEC1, 6 in DEC2, and 3 in DEC4. The logic connected to DEC3 is shown in fig. 7.

DEC3 also decodes the commands necessary to enter the equator crossing longitude and time since crossing from the external BCD switches. As seen in **fig. 7**, the switches and outputs from the timers are OR-wired and used to feed a 7445 BCD-to-decimal

table 3. DEC output functions. In DEC 2, only the underlined functions are used in the program.

binary	DEC 1 decoded function	binary	DEC 2 decoded function		binary	DEC 3 decoded function	binary	DEC 4 decoded function
000000	not used	010000	not used		1000000	not used	110000	not used
000001	1	010001	SHIFT	(DPS)	100001	EQX 100	110001	sine
000010	2	010010	CLX	(clr)	100010	EQX 10	110010	cosine
000011	3	010011	ENT	(sci)	100011	EQX 1	110011	tangent
000100	4	010100	STO	(INV)	100100	EQX .1	110100	not used
000101	5	010101	RCL	(HYP)	100101	EQX .01	110101	not used
000110	6	010110	EE	1/x	100110	(T) 10 (minutes)	110110	not used
000111	7	010111	ln	(log)	100111	(T) 1 (minutes)	110111	not used
001000	8	011000	sigma +	(x,s)	101000	Go Sub 1	111000	not used
001001	9	011001	+ / −	(x!)	101001	Go Sub 2	111001	not used
001010	0	011010	x↔y	(%)	101010	Go Sub 3	111010	not used
001011	D.P.	011011	roll x	(delta%)	101011	not used	111011	not used
001100	+ (add)	011100	y^x	(sqr x)	101100	not used	111100	not used
001101	− (sub)	0111101	not used	not used	101101	not used	111101	not used
001110	x (mult)	0111110	not used	not used	101110	not used	111110	not used
001111	÷ (div)	0111111	not used	not used	101111	not used	111111	not used

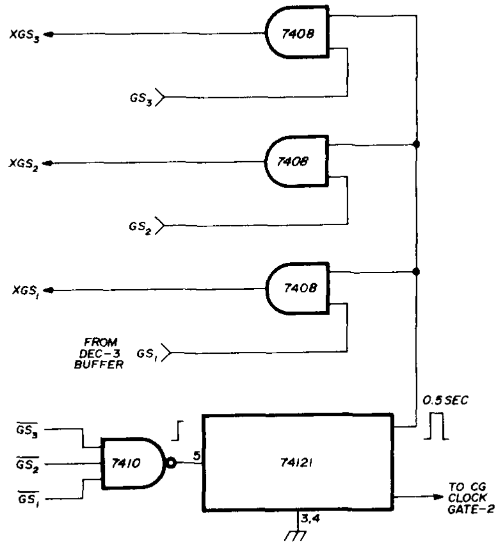

fig. 5. Details of the GoSub routine logic. The output pulse, in conjunction with the digit strobe, is used to enter the output data into the 7475 latches.

decoder. The output from the decoder also drives reed relays connected across the calculator keys.

Fig. 6 shows digital information indicating the actual azimuth and elevation of my antennas. In addition to being applied to the 7485, 4-bit comparators, the information is applied to BCD-to-7 segment decoders and readouts. Also, **fig. 7** indicates where decoders could be connected to readout the time since equator crossing. With all the information converted for readouts, I have a panel that shows displays of: azimuth (beam), azimuth (calculated), elevation (beam), elevation (calculated), distance M (calculated), distance M (for later use), time, and sequencer location. The sequencer location was included as a troubleshooting aid should the program ever stop.

Limit switches have been included in my system to stop the antenna from going beyond the prescribed limits. If you run the program only during valid pass times, the program should never produce invalid commands. But, should this ever be a problem, the limit switches will prevent major damage. You can readily see from the readouts where the problems are if they occur.

Trigonometric functions near 0 or 90 degrees, and numbers which result in zero denominators *can* give the program fits, but there just isn't any easy way around this. I haven't found it to be a problem, however, except on way out, very short passes. A final note on PROM programming: in steps 059-063 and

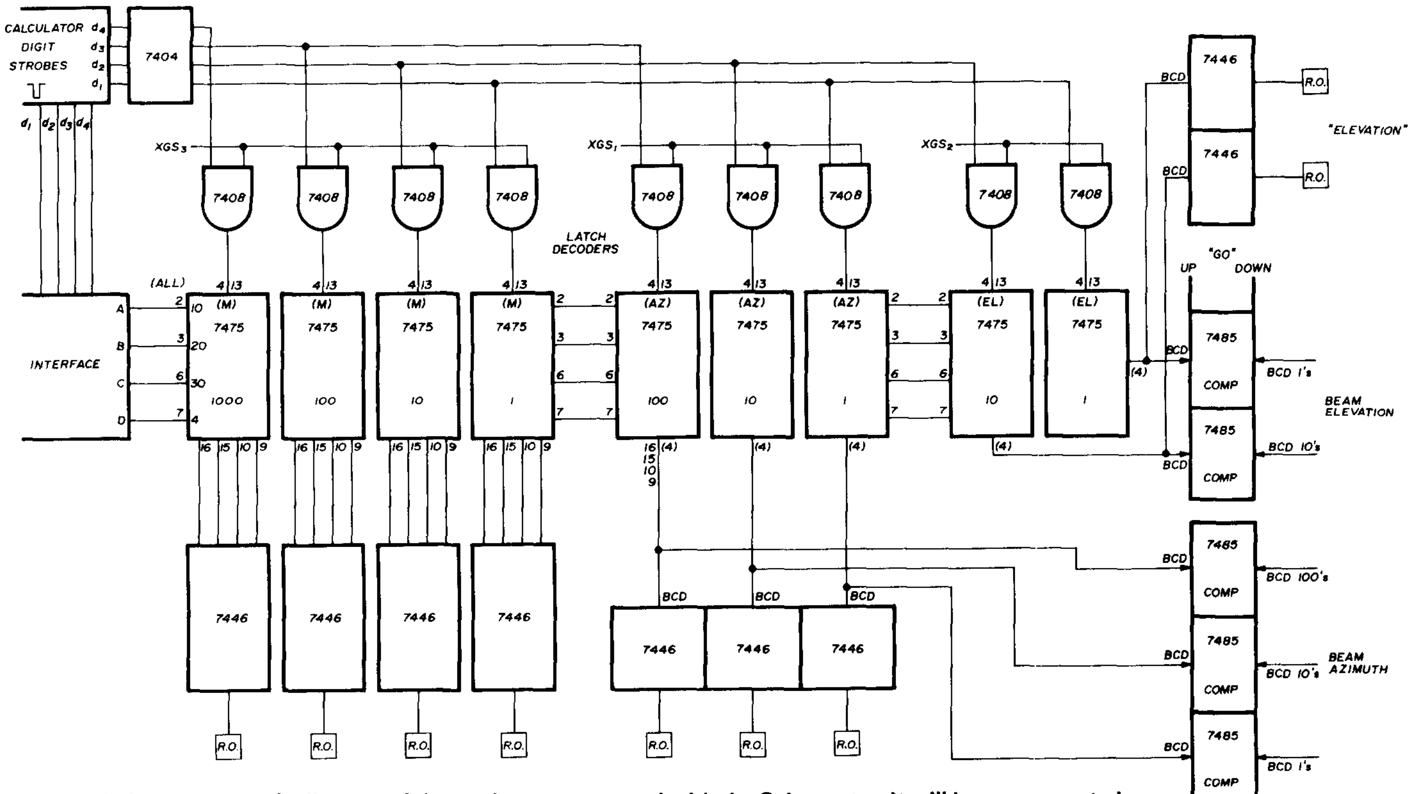

fig. 6. Quasi-schematic diagram of the readout system used with the Calcu-puter. It will be necessary to insert an interface device between the 7-segment information and the BCD format required by the 7475s. The author uses another 1702A EPROM programmed to do the necessary conversion. An alternate method would be to use the National 74C915 to convert the data. Though not shown in this diagram, the author has also connected decoders and readouts to indicate the actual antenna position. The 7404 buffers between the calculator and the latches may have to be changed depending upon the type of strobe coming from the calculator.

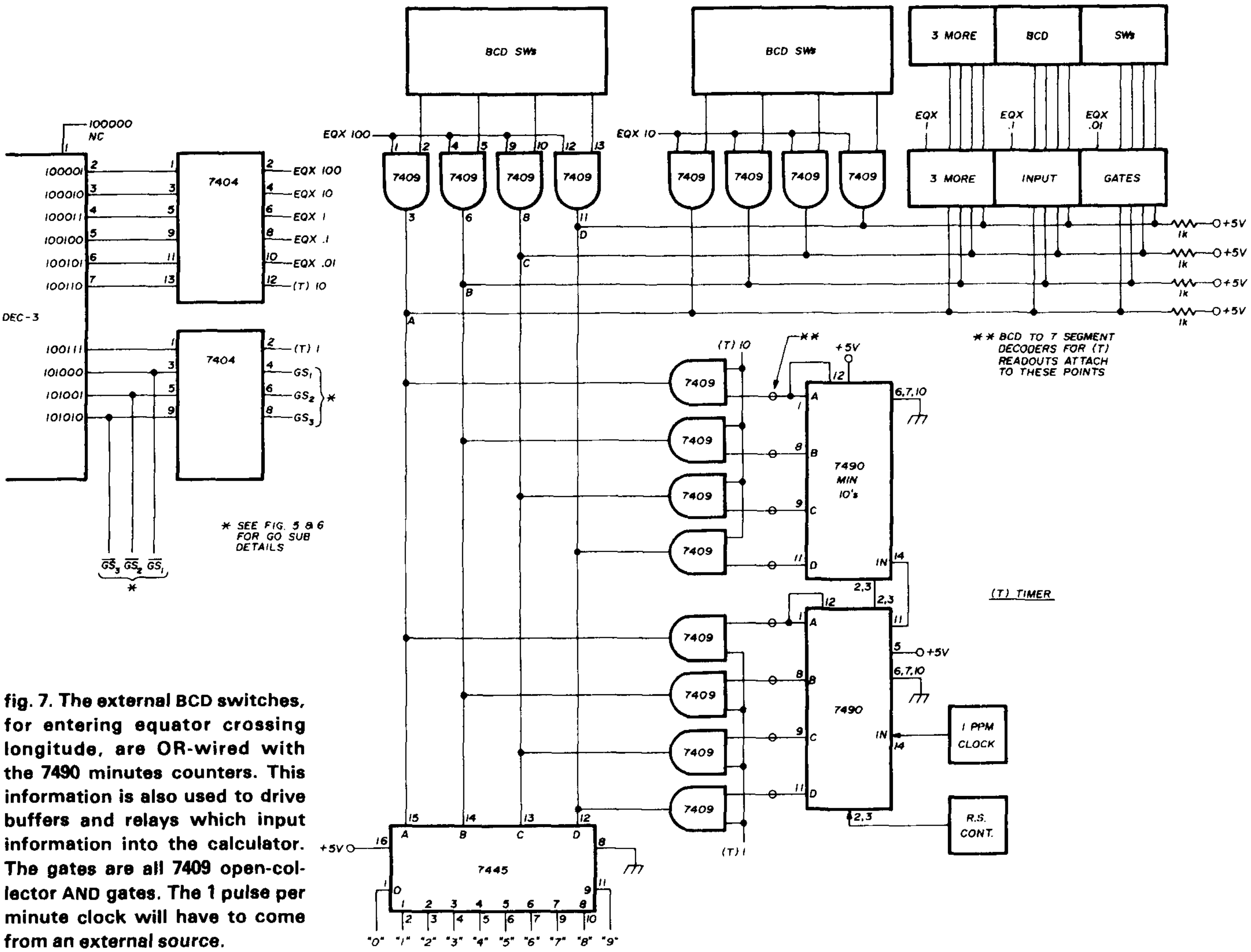

fig. 7. The external BCD switches, for entering equator crossing longitude, are OR-wired with the 7490 minutes counters. This information is also used to drive buffers and relays which input information into the calculator. The gates are all 7409 open-collector AND gates. The 1 pulse per minute clock will have to come from an external source.

087-092 be sure to enter the latitude and longitude for *your* location. This will be retained as permanent information in the PROM.

concluding comments

The primary message of this article has been to show you that a complete computer/microprocessor is not required to do simple math problems. The PROM is in a sense a simple BASIC language like no other. It has automated a calculator, providing both for inputing and outputing of data, much in the same way as a full-scale computer.

I did write a program in algebraic notation instead of RPN, but quickly discarded it when I couldn't find an inexpensive calculator with enough onboard memory. Lacking this capability meant dumping out the interim answers, performing more calculations, and retrieving the interim answers before the final numbers could be outputed. It generally amounted to a lot more hardware, fast approaching a full-blown computer, a mess that I wanted to avoid from the beginning. The APF 55 calculator I finally used was provided by a friend because some of the digit segments would not light. It was about as cheap, and definitely quicker, for him to buy a new calculator and give me his remains! Shop around because the price on some of the very sophisticated units is getting ridiculously low. For that matter, one of the many calculator shops around these days might part with some of their damaged returns, for the right price.

For anyone wishing additional information, a self-addressed, stamped, envelope will bring a quick reply; and any comments on improvements to the system will be welcomed.

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1. Bob Henson, WB0JHS, "Computerized Satellite Tracking," *73*, February, 1977, page 72.
2. Bruce McNair, N2YK and Glenn Williman, N2GW, "Digital Keyboard Entry Systems," *ham radio*, September, 1978, page 92.
3. David Brown, W9CGI, "Introducing Autotrak," *73*, July, 1977, page 46.

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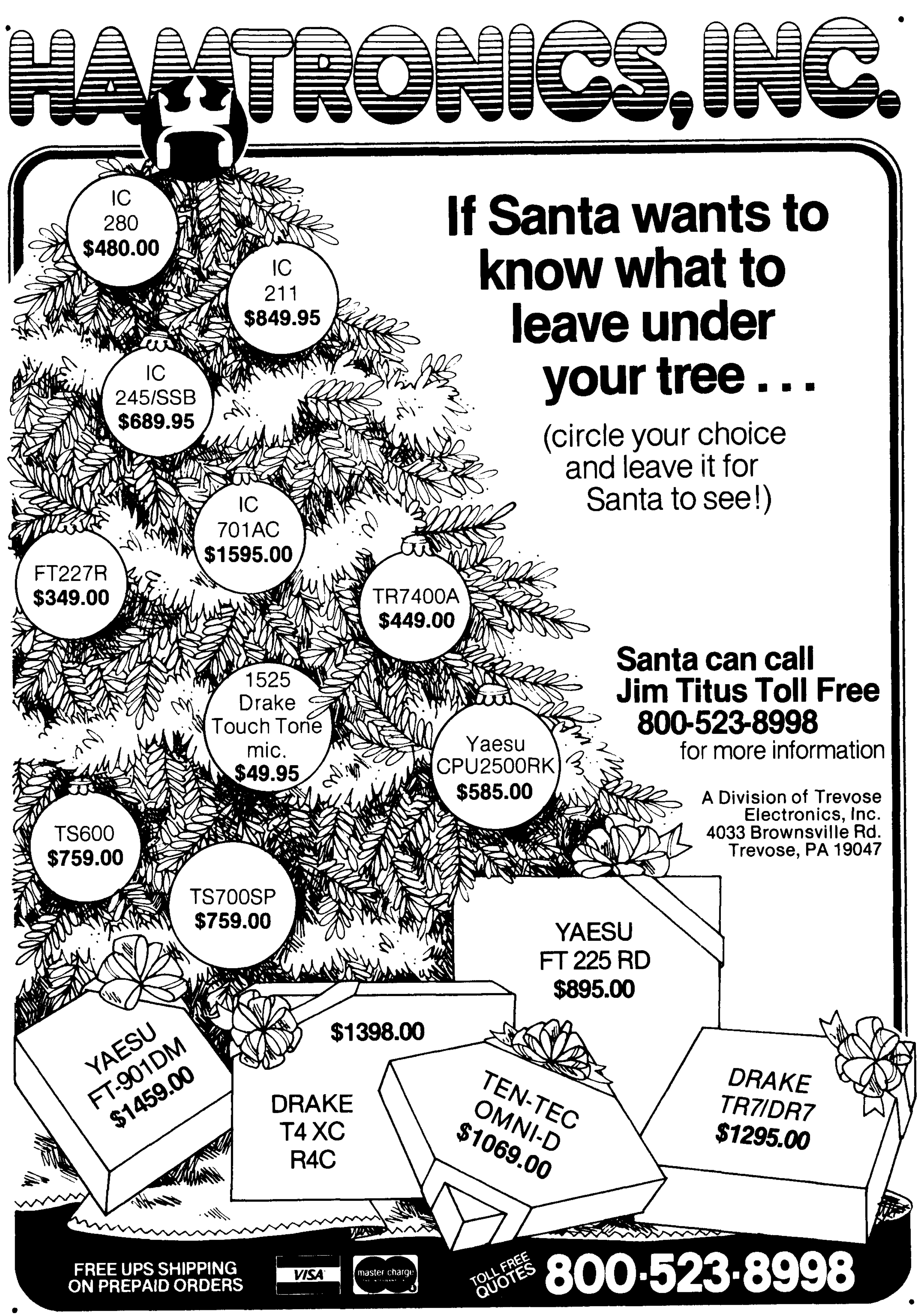
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simple video display

Two projects to get you started in building a video display unit using readily available devices

There has been much recent interest in video display units. They can be used as part of a video typewriter, for putting up displays on ATV and SSTV, and for decoding RTTY and Morse off the air. Most of the displays are complex and expensive. Even the available kits aren't suitable for those who haven't had much experience. For normal use you don't need whole screens full of characters, and the simple 32-character single-line display described here is an excellent beginning for those who would like to play around with an inexpensive video display unit.

description

The heart of the unit is the Fairchild 3258 dot-matrix character generator IC. Externally it's a 16-pin package (I hate to think what's inside it!), which accepts ASCII inputs and produces 64 characters on a 5 x 7 dot matrix. Apart from the inputs and outputs, the only other signal connections to the chip are inputs to a clock and a master reset. The chip has an internal clock and addressing system. After the master reset input operates and goes high, the information representing the first row of the character is available after the first clock pulse. Subsequent clock pulses select the next six rows in turn; after that, the outputs are clamped high and the character generator stops until another master reset pulse appears. Thus, if the character-generator clock is pulsed at line frequency, the character will appear on the screen.

experimental system: one character 32 times

Fig. 1 shows the logic diagram of the display unit. Only nine integrated circuits, including the character generator, are required. There is no reference crystal or dividing network. I used a monitor from a non-composite camera and monitor combination and simply fed the horizontal and vertical sync pulses into the VDU. If you wish to use a regular TV set, it's quite easy to add two 555 timers to provide horizontal and vertical sync pulses. **Fig. 2** shows the connections for the monitor, which would be typical, and **fig. 3** shows the circuit for the 555 timers required.

system operation

The second half of the 74123 feeds a gate, which feeds a second gate, which in turn feeds back into the 74123. This action sets up an oscillating circuit whose frequency is determined by the 5k pot in the +5-volt line. This frequency is used to step the 74195 shift registers and provides the basic video signal. The character-generator output is loaded into the 74195 shift registers then clocked out in a serial mode at the VIDEO OUT terminals. The **J** and $\overline{\mathbf{K}}$ inputs of the first shift register cause highs to be entered as the data from the character generator is shifted along. Finally, when the six outputs to the 9007 gate are all high, a low is sent on the MOD 7 counter line, which reloads the shift registers and clocks the **horizontal character count** 7493 ICs.

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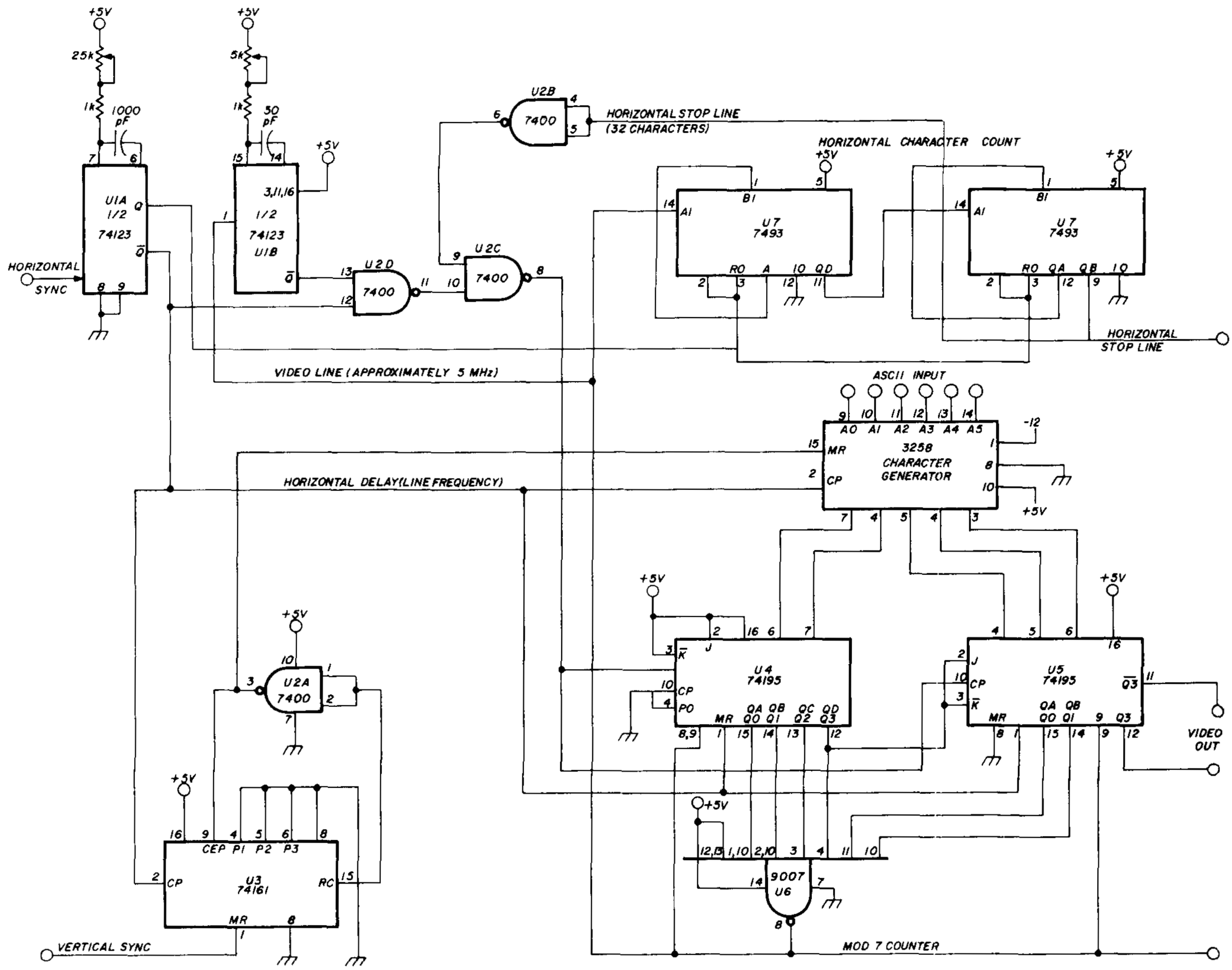

fig. 1. Logic diagram of the video display unit (VDU). Only nine ICs, including the character generator, are used.

When this action occurs 32 times, the 7493 provides an output to the HORIZONTAL STOP line, which inhibits the 5-MHz oscillator and stops the sequence.

This action occupies about two-thirds of a single horizontal line. When the end of the TV line is reached, a horizontal sync pulse operates the first half of the 74123, resetting the 7493 counters, providing a clock pulse to the character generator. This pulse acts to output the information for the next line and resets the 74195 shift registers. The sequence then repeats for the next line.

The character generator automatically blanks out after a complete row of characters has been sent, and if it's required to have more than one row, the 9316 will count the rows and reset the character generator. A vertical sync pulse resets the 9316, so the information is always at the same position on successive frames.

There are no critical adjustments in this circuit. The 25k pot in the first half of the 74123 positions the first character on the left-hand side of the screen, and the 5k pot in the second half opens up or closes the 32-character-length display so it can be spaced evenly across the screen.

The logic shown in **fig. 1** will produce a display of one character repeated 32 times across the screen. The character will be determined by the ASCII input to the character generator. For test purposes you can apply a combination of 5-volt and ground inputs as required. This can be treated as a project in itself, so that those who want to take a bit at a time can do this, then go on to the second half of the project.

32 different characters

The logic for the second half of the project is shown in **fig. 4** and again is quite simple. Here, the main device is a Fairchild 3349 hex 32-bit static shift register. Like the character generator, the 3349 has a

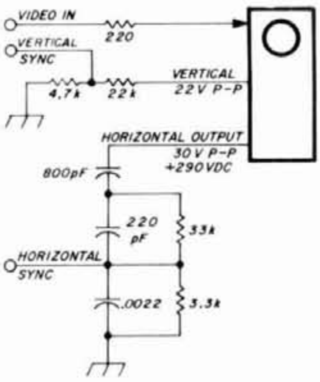

fig. 2. Connections for a typical monitor (Toshiba TMC2AX).

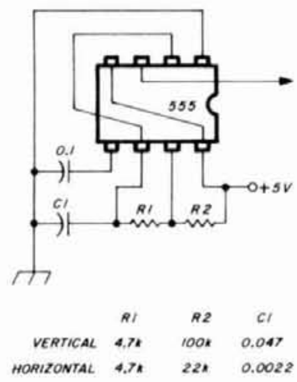

fig. 3. Circuit using the NE-555 timer to provide horizontal and vertical sync pulses if a regular TV set is used as a monitor.

deceptively simple 16-pin package; and again, there is a complex integrated circuit inside.

Only two signal controls are needed: a clock input and a load/recirculate input. The clock pulse steps the 32 bits in each of the shift registers, which recirculate until the LOAD/RECIRCULATE input goes low. Then new data is accepted and the data at the other end is lost. When the LOAD/RECIRCULATE input goes high again, the 32 bits in each of the six shift registers at that time resume recirculating. To obtain 32 different characters across the screen, the shift registers must present the six new bits to the character generator each time the Mod 7 counter operates so that the shift-register clock is fed from the Mod 7 counter. This action would produce 32 characters, but they would be random characters that happened to come up when the display was first switched on. So we must have some way of putting in the characters we want. This is done by simultaneously presenting the required ASCII code to the shift-register inputs and applying a negative key pulse to the set input of a flip-flop. This action sets the flip-flop output high and puts a high on the data input of a second flip-flop (both halves of a 7474). A pulse from the HORIZONTAL STOP line (at the end of the display of the 32 characters) clocks the second flip-flop. This allows a low to be put on the load input and also operates a gate, allowing an extra clock pulse (from the horizontal sync) to clock the new data into the shift registers. Then, at the end of the horizontal sync pulse, the flip-flops are reset and the 31 old characters and the one new one recirculate until another character is entered.

The two 74121s merely give a controlled-length

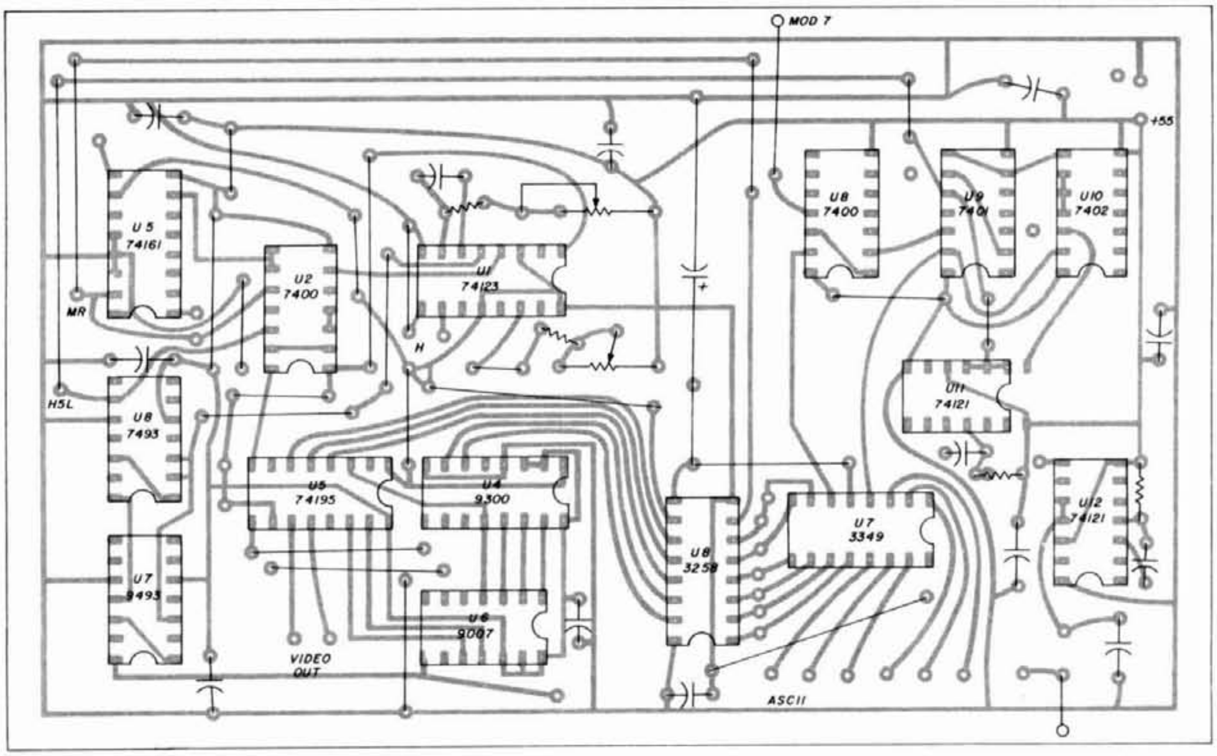

fig. 5. Full-size PC-board layout for both sections of the display unit. It has been tested by the author and works perfectly.

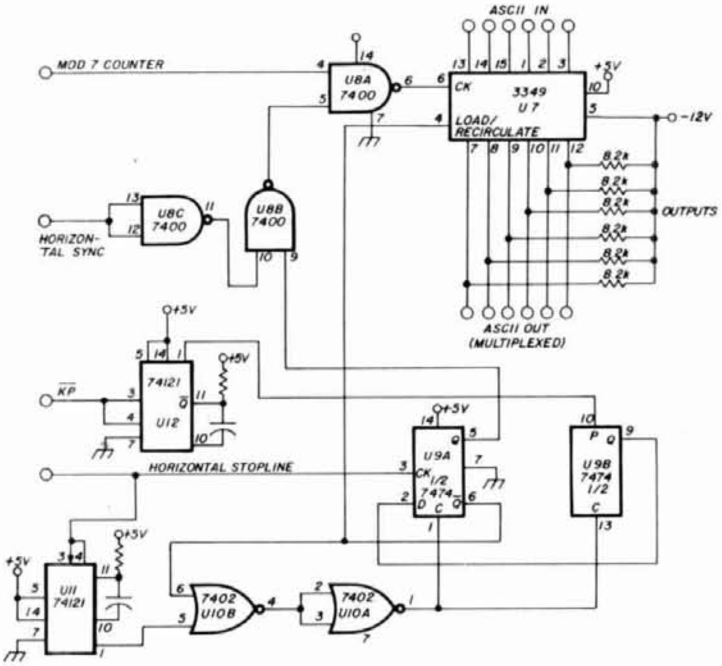

fig. 4. Video-display logic for producing 32 different on-line characters. The main device is a Fairchild 3349 hex 32-bit static shift register.

pulse and could be replaced with a resistor-capacitor combination, but they were used because the pulse length is more controllable. Apart from the shift register and flip-flops, the only other ICs are a couple of normal gates. The only other point worth mentioning is that the shift register outputs require external resistors (8.2k) from each output to the −12-volt supply.

construction

Two simple circuit boards about 3 inches (76mm) square will accommodate the entire system, or it can be built on a slightly larger board (**fig. 5**). It's a good idea to use sockets for the character generator and the shift register. Sockets for the other devices are a matter of personal preference.

When testing the circuit be very careful not to let the −12-volt supply get into any of the +5 volt TTL devices — it can have disastrous results!

final remarks

This project will give a beginner in this area an insight into the principles of VDUs and provide an excellent starting point for developing something more complex. To keep the project as simple as possible no attempt has been made to eliminate additional lines, so there will be several identical lines of 32 characters across the screen. In practice, they help rather than hinder reading the characters.

ham radio

updating the Collins 32S-1

Never being satisfied with the *status quo* when it comes to any radio equipment I've ever owned, I eventually succumbed to the urge to modify my recently acquired Collins 32S-1 transmitter. The modifications described here include the following:

1. BFO generation of the CW carrier
2. Voltage regulation of the PTO and HFO
3. Control of the keyed wave shape
4. A spotting switch (CW CAL)
5. The ability to monitor the final-amplifier plate (cathode) currents individually
6. Alterations to the tone oscillator

The modifications were made to bring the performance of the 32S-1 up to the standards of its successor, the 32S-3, without incurring an expenditure of some $300-$400 in the process.

Table 1 identifies the components involved in the modifications discussed here. Schematics and parts lists should be changed accordingly to reflect these changes, since removed components will have their identities transferred to newly installed pieces that correlate with those used in the 32S-3.

BFO CW generation

The 32S-1 generates its CW carrier with a tone fed from the tone oscillator through the mechanical filter (much like whistling into the mike or feeding AFSK into the mike jack on RTTY). The frequency of the tone used in the 32S-1 was chosen specifically so that its second harmonic falls well outside the mechanical-filter passband. However a weak residual signal still exists, and it has been heard on occasion at some distance.

The 32S-3, uses the BFO signal to generate the CW carrier, eliminating this residual signal. The resultant on-the-air signal is much cleaner and sounds much more like a true CW signal when compared with that of the 32S-1.

Installing this feature requires extra switching capabilities, which must be performed by the EMISSION switch, S8. The 32S-1 has four wafers on this switch, while the 32S-3 has five. Here are some ways in which this additional switching may be handled; a separate 4 PDT toggle switch may be used; S8 may be entirely replaced; or the existing switch may be disassembled and a new index and wafer added. Although the first possibility was initially pursued, I found it to be inconvenient. The most satisfactory arrangement was to replace the index assembly and add an additional wafer to S8.

The MIC GAIN pot and switch must also be replaced with a new unit using two pots commonly controlled and switch S14. The additional pot controls the cathode bias (CW DRIVE) on the rf amplifier, V6. Both parts are available from Collins; the switch is part no. CPN 259-1628-000 and the dual pot and switch is part no. CPN 376-2648-0000.

First, replace the existing MIC GAIN pot with the new dual unit. Note that space is at a premium, and the possibility of a shorted terminal strip lug exists next to V12. To avoid this, mount a two-lug terminal strip on the opposite side of the crystal board and secure it with the self-tapping screw that holds another two-lug strip. Remove the B+ ends of R60, L20, and the B+ feed wire (green/white) from their original location. Attach them to the new terminal strip. The now empty lug may be bent over to clear the pot and switch R8/S14.

Mount a single-lug terminal strip under the hardware securing the two ground lugs between V13 and V4. Lift C20 (0.01 μF) from ground and connect it to the strip. Route a length of RG-174/U cable from this junction to the vicinity of S8. Lift R39 (V6, pin 7) from ground and connect that end to a single-lug strip that has its ground lug straightened and soldered to the ground shield/barrier across V6. From this same point, run a wire to R8B and install a new R71 (68k/2W) between this lug and the terminal lug near V5 where R29 and R30 (4.7k/2W) connect to the +275-volt line (red/white wire).

From R8B run another wire to S8-B lugs 9 and 10, which are then connected in parallel. In the 32S-1

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these two lugs are empty, as are those on the wafer to which R87 (470 ohms) is attached. Connect the empty lug of R8B to ground.

Connect a 33-ohm resistor (new R70) to V2A pin 9. Remove the BFO input cable. At this time S8 should be modified or replaced. Assuming the index and wafer are to be replaced and added, remove the switch and thread some bare wire through the rivet holes (which secure the switch contacts on wafers 1 through 4) at two points 180 degrees apart to prevent the spacers from separating from the wafers. Then the existing index may be removed, replaced, and the 5th wafer added with little effort.

Wire the switch as shown in **fig. 1**. Run a wire from S8B lug 11 (presently empty) to V10 pin 1 to prevent premature VOX relay dropout on CW.

ALC modification

Unlike the 32S-1, the 32S-3 does not use ALC in the CW position. During CW, switch selection S8G-5 grounds the midpoint of ALC capacitors C83 and C142. This change may be added to the modified

table 1. Component identification for the 32S-1 mods described in the text.

32S-3 part no.	original 32S-1 part no.	value	modified 32S-1 part no.	value	location
C81	not used		C81	0.005 μF	second mixer
C115	C115	0.01 μF	C115	0.33 μF	keying circuit
R17	R17	33k/1W	R17	5k/10W	voltage regulator
R70	R70	470k/1/2W	R70	33 ohm/1/2W	V2A
R71	R71	470k/1/2W	R71	68k/2W	B+

32S-1 by simply adding a jumper wire from S8G-1 and -2 to S8G-5 (**fig. 2**). Now, during CW operation, the GRID CURRENT position (instead of ALC) is monitored, and the MIC GAIN control is adjusted to obtain a grid current reading of 1 to 2 dB on the meter while sending a series of dots.

keying circuit and CW calibrate

The 32S-3 keying circuit provides some manual control of the keyed wave shape, **fig. 3**. The spotting feature (CW CAL) may be installed coincidentally. The CW CAL function switch should be front-panel mounted for ease of operation. The KEY SHAPE control, R123, may be located under the lid of the 32S-1 exciter on the bracket containing the VOX controls, or a separate bracket can be made and attached to the power-amplifier cage with self-tapping screws. Most of the other components are mounted on the terminal strips from which the 32S-1 keying circuit components will be removed. The addition of a single three-lug terminal strip (center ground) between K1 and V14 ensures that all components are securely mounted.

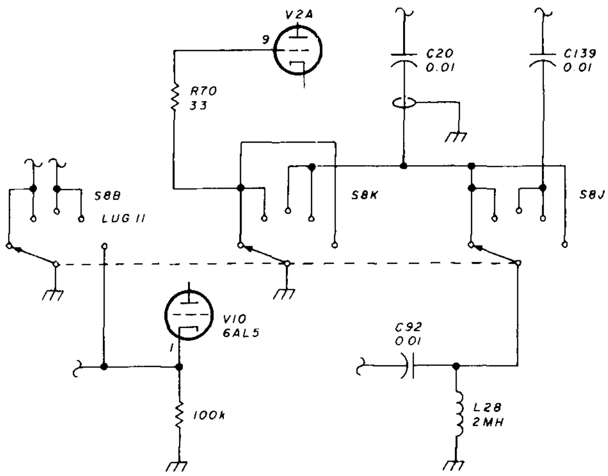

fig. 1. Schematic of the BFO generated CW showing modification of switch S8 to eliminate the weak residual signal in the 32S-1 when in the CW mode.

R70, R71, and R72 may be removed from the terminal strips at the bottom left of the chassis and R125 mounted in place of R72; R126 in place of R71; and R124 in place of R70. Remove relay K1's lead and mount it onto the newly installed terminal strip.

Instead of using the multiple-leaf switch and 250k pot arrangement of the 32S-3 for the CW CAL function, a fixed resistor and three-pole rotary switch were used (**fig. 4**). The rotary (or toggle) switch has a more positive action and doesn't require constant depression to activate the desired function. A value of 68k resulted in a satisfactory over-all spotting level and this resistor was secured to the two innermost lugs of a 5-lug (center-ground) terminal strip mounted with its ground lug soldered to the ground lug of the strip behind K1 and at right angles to it. (The other lugs will be used in the regulated voltage modification.)

Mount the 3PDT switch (S13) on the front panel between the FREQUENCY CONTROL and MIC GAIN shafts. Center the holes 87 mm (3-7/16 inches) from the top of the panel. If done carefully it will appear to have been factory installed.

For ease of wiring and installation I recommend that the FREQUENCY CONTROL switch be temporarily removed. Unsolder and tie back the green/white

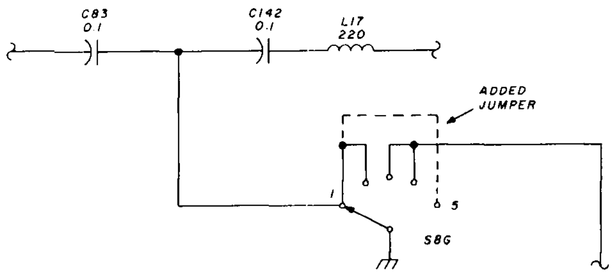

fig. 2. A jumper wire is added to switch S8-G to ground ALC voltage during CW operation.

wire at S9E-1. Wire the remaining circuit according to **fig. 4**.

In operation the transmitter must be properly tuned for CW operation for the CW CAL function to be enabled; it will not work on ssb.

The KEY SHAPE control (R123) should be adjusted to eliminate key clicks created by the rapid rise of the keyed signal. The effect of this control will be fully appreciated when the transmitted signal is monitored on an oscilloscope. The control should be adjusted to round the leading edge of the waveshape slightly.

Additional shaping of the waveform on the trailing edge may be accomplished by adding capacitance in two places: between the key line to ground and between the junction of R33/R37 and ground in the

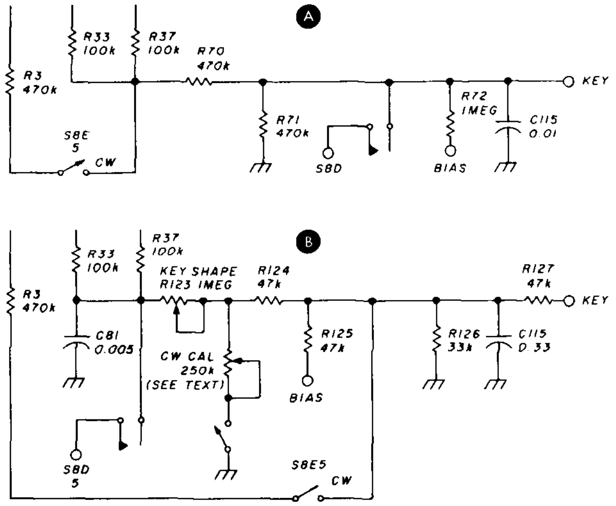

fig. 3. Keying circuits of the 32S-1 (A) and 32S-3 (B). By modifying the 32S-1 as described, you can have control of the keyed waveform within limits. The CW CAL feature is a handy addition. It won't work on ssb, however.

first mixer, V5. Some experimentation should provide a wave with the desired characteristics, with values of 0.025 μF (C115-A) and 0.005 μF (C81) being a good starting point in their respective positions. See **fig. 5**.

A difference will be noticed between on-the-air signals when using a transistor-output keyer versus a bug or relay-output keyer; the transistor provides a softer signal and you might use considerably more key-line capacitance with a bug or relay-output keyer, depending on personal preference and speed. Too much capacitance at high speeds tends to slur the code elements.

voltage regulation

In the 32S-3, the 6AL5 ALC rectifier was deleted and solid-state devices used in the ALC circuit. This

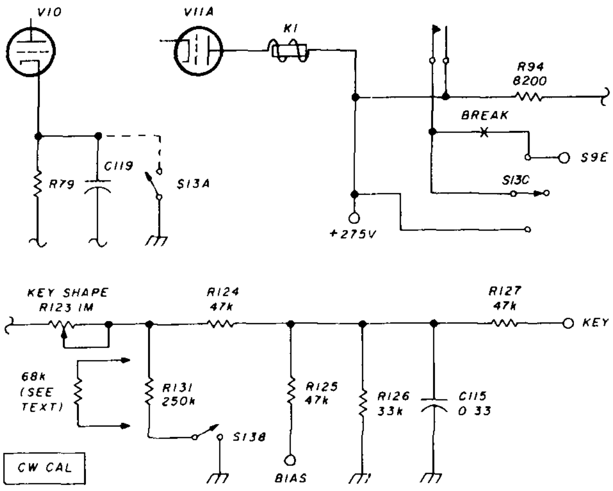

fig. 4. Modifications to the CW CAL circuit. A fixed resistor and a 3-pole rotary switch provide more positive action. It isn't necessary to hold down the switch to activate the desired function.

freed socket V13, which was used to hold an OA2 tube to supply the regulated voltage for the oscillators. I found it simpler to use a 140-volt, 10-watt zener (1N3010A) for the regulator. They are inexpensive and eliminate the need to free V-13's socket, with the problems of rewiring the ALC circuit and finding space for more parts.

An advantage of the zener is its ease of mounting. Mount CR9 (**fig. 6**) on the perforated wall of the bottom side of the power-amplifier cage by enlarging one of the holes to accept the 10-32 threaded stud of CR9. Mount a dropping resistor (new R17, 5k/10W) on the terminal strip installed previously to the rear of K1. (Note: The original R17 must be removed according to the following steps.)

A convenient source of +275 volts is the terminal of C137 on the PA-cage wall; it has the 100-ohm/1/2-W resistor attached.

Modify the PTO and HFO circuit as follows. Remove the original R17 (33k/1W) and substitute

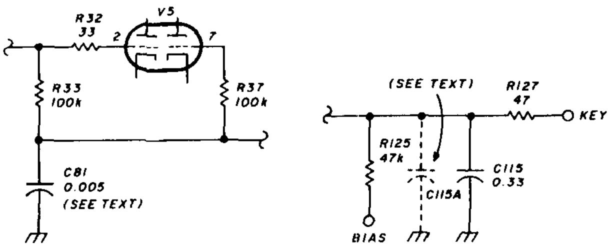

fig. 5. Improved keying wave is provided by this change. The added capacitances improve shaping of the signal trailing edge. Capacitances C115A and C81 (respectively 0.25 μF and 0.005 μF) are good starting points. Some experimentation might be needed to provide desired waveform characteristics.

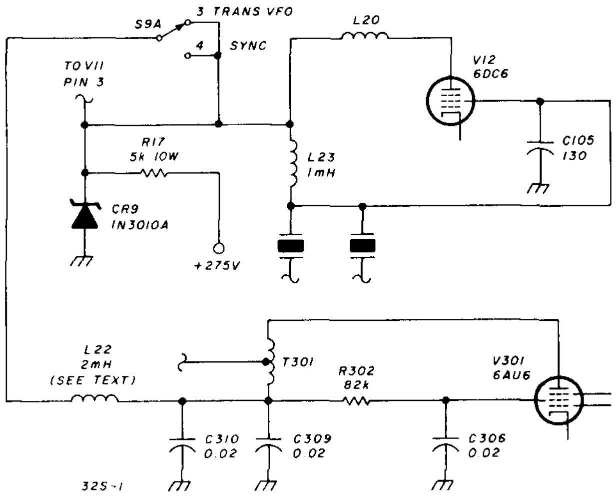

fig. 6. Regulated-voltage modifications. A 140-volt, 10-watt zener replaces the old OA2 regulator. This change is even simpler than that in the 32S-3 (see text).

L22 (2-mH). R17 is located close to C57 and the shield can. Run a wire from CR9 past the crystal board and up through the grommet to S9. At S9, locate the red/white/green/blue wire that connects to L22's B+ end. Cut the black jumper connecting the two S9 wafers (+275 volts) and attach the +140-volt line to the commoned lugs, 3 and 4, TRANS VFO and SYNC (**fig. 6**).

One of the two green/white wires on S9's rear wafer supplies +275 volts to the HFO, V12. Locate this wire, disconnect it at S9 rear, and move it to the +140-volt line on lugs 3 and 4. Disconnect R60 (47k) completely. Install L23 (1 mH) in its place. This completes this modification.

tone-oscillator changes

Both before and after the modifications described, an unwanted high-frequency oscillation was audible. I found it necessary to add 0.1 μF of capacitance between V11 screen and ground in parallel with C107.

Since the tone oscillator no longer supplies the on-the-air CW signal, its frequency may be altered to provide a more pleasant monitoring note. This note is purely a matter of personal preference, so some experimentation may be necessary. In my case, a 100-pF mica capacitor was paralleled with C110.

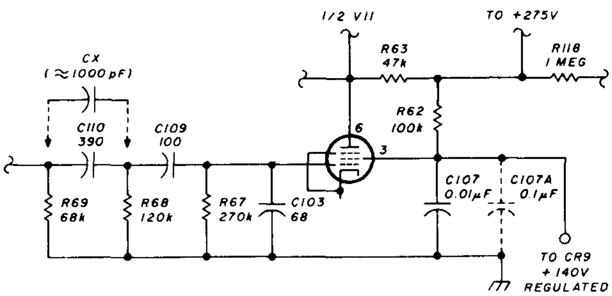

fig. 7. Tone oscillator changes. This change eliminates an unwanted high-frequency oscillation in the output. Capacitor CX in parallel with C110 provided a more pleasing monitoring note.

separate plate-current monitoring of the power amp

Unsolder R52 (1k). (Note: This value may differ from unit to unit.) R52 is attached to the copper strap joining the cathode pins of the two 6146s. Cut and remove the strap from between the tubes. Attach a length of hookup wire to each of the pins from which the strap was removed and route them toward the perforated wall of the PA cage. Mount a 4-lug terminal strip inside the enclosure toward the rear of the chassis with 4-40 (M3) hardware and wire as shown in **fig. 8**. Mount the DPDT switch, S15, directly beneath the meter. For ease of access the meter should be removed before drilling the mounting hole. Use a miniature toggle switch in this location, which is almost unnoticeable.

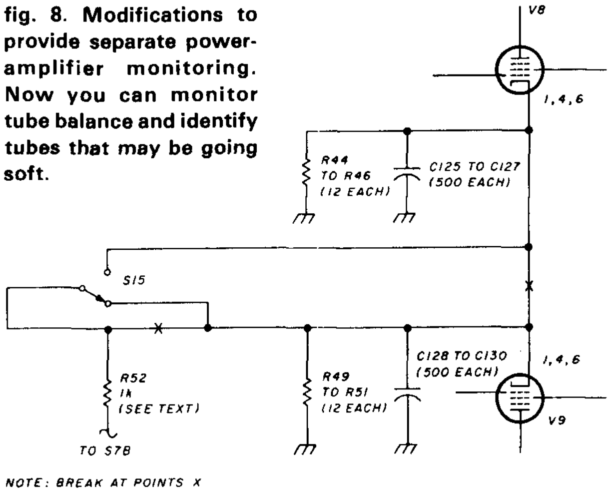

fig. 8. Modifications to provide separate power-amplifier monitoring. Now you can monitor tube balance and identify tubes that may be going soft.

No interpolation of the readings is necessary since the cathode voltage/resistance ratios are unchanged. Tube balance, which is necessary in all parallel-tube amplifiers, is readily observed, and a soft tube may be easily spotted. The cathode currents of the individual tubes should track within +10 per cent to satisfy a balanced condition.

closing remarks

The incorporation of these mods into the Collins 32S-1 provided performance that rivals that of the more costly 32S-3. It's given a new lease on life to a veteran of some 18 years and has saved a couple of *hundred dollars in the process!*

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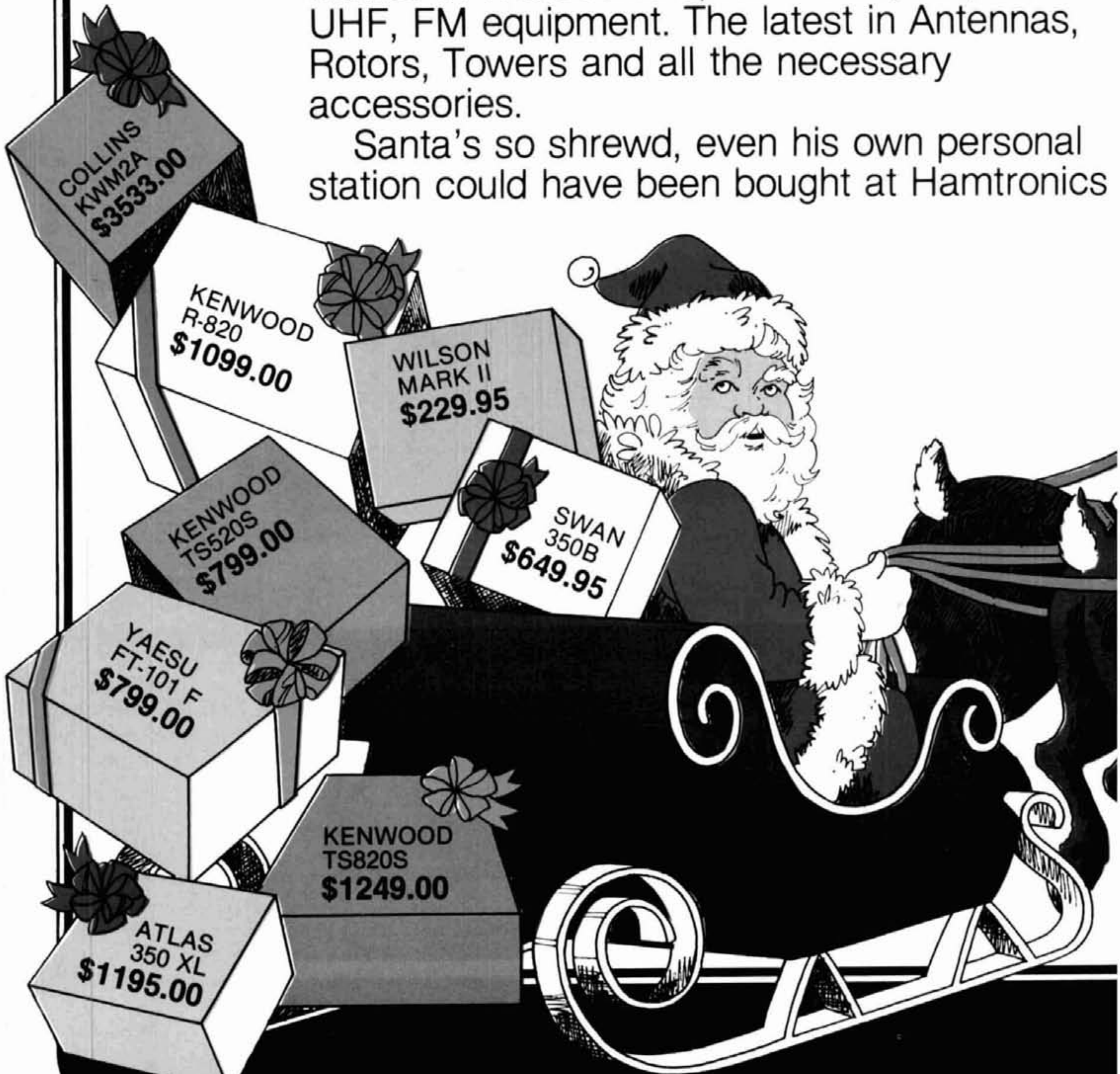

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top-loaded delta loop antenna

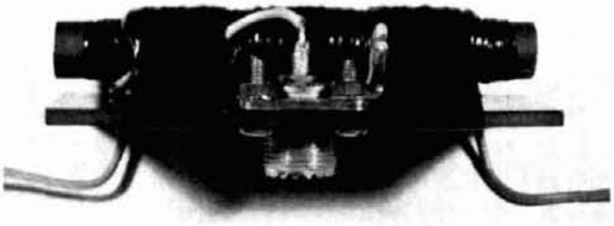
Design and construction of an efficient, low-frequency, vertically polarized antenna using wire elements

The vertically polarized full-wave loop has emerged as a popular antenna on the low-frequency bands. The most common form of this antenna is the triangular (delta) loop[1,2] with one of its vertices pointing skyward. Such an antenna can be suspended from a single point located on a tower or a tree.

The delta loop antenna is an interesting cousin of the popular inverted-V dipole. It has been around for quite a while and yet provides some pleasant surprises. For those interested in tracing its background I have provided references 1 and 2. Reference 1 is particularly informative and provides polar diagrams of the delta radiation pattern in three planes together with supporting mathematics. These references are available in most of the libraries in large cities.
Editor, W6NIF

On the 80- and 160-meter bands, height limitations can reduce the effectiveness of the delta loop. This article describes a method for reducing this problem by means of an easily implemented loading procedure. The case of a support height of 20 meters (65 feet) for an 80-meter antenna is shown in **fig. 1**. An interesting aspect of this comparison is that the top-loaded delta loop **fig. 1B** (TLDL) has more gain than a full delta loop. Experience since the end of 1976 at W1DTV has been that the antenna performs as well as an inverted V for short-range contacts and provides one to two S units better performance for DX contacts. In this article, I discuss the evolution of the TLDL and provide detailed design information for an 80-meter TLDL.

Two kinds of vertically polarized antennas are in



Homebrew matching transformer for the top-loaded delta loop antenna.

By Frank J. Witt, W1DTV, 20 Chatham Road, Andover, Massachusetts 01810

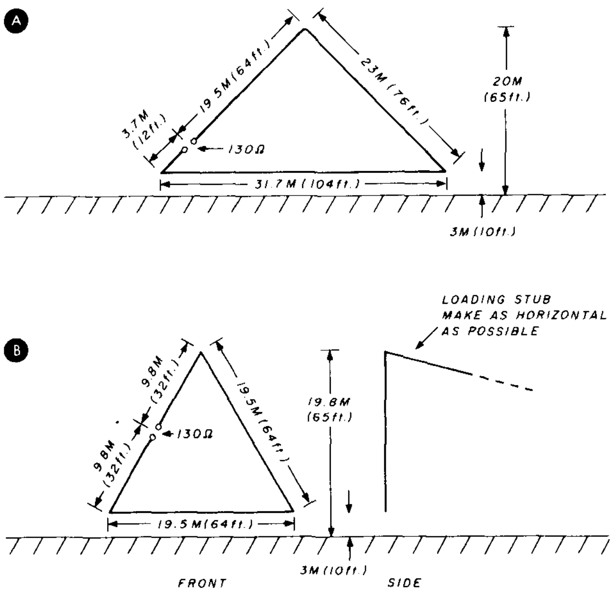


fig. 1. An 80-meter delta-loop antenna with apex at 20 meters (65 feet). ***Sketch A shows the classic delta loop for*** **3.825 MHz;** ***B shows a top-loaded delta loop for the same*** **resonant frequency. Loading-stub dimensions are discussed** ***in the text.***

common use on the low-frequency bands. One type is suspended above ground and fed directly; the other is erected from ground level and excitation occurs between ground or a simulated ground plane and the antenna. Both antennas would benefit from a highly conductive ground; but in the latter case, since ground resistance appears in series with the antenna at the drive point, efficiency is highly dependent on ground conditions. Therefore, the more successful monopole installations are those that use many radials. The TLDL is not fed against ground and hence ground plays only the role of a reflector. This is also true of full delta loops and sloping dipoles. Experience has shown that impressive performance may be obtained with such antennas without an elaborate system of radials.

evolution of the top-loaded delta loop

The signal at a distant point from a part of a transmitting antenna is proportional to the current in that part of the antenna. For a half-wave dipole, for instance, maximum radiation is received from the center of the dipole, where the current is greatest. The radiated contribution from the ends of the antenna is negligible.

The TLDL concept resulted from a recognition of the fact that for a conventional, vertically polarized delta loop, much of the antenna where high currents exist is horizontal and near ground. The objective of the TLDL design is to get these parts of the antenna away from ground and at least partly vertically oriented to increase antenna gain. **Fig. 2A** shows a typical vertically polarized conventional delta loop designed for 3.825 MHz. Actually, this antenna can only be said to be mostly vertically polarized because of the position of the feed point. True vertical polarization (in a direction perpendicular to the plane of the loop, *i.e.*, the direction of maximum gain) is obtained when the feed point is one-quarter wavelength away from the peak of the triangle as shown in **fig. 2B**. You can see that the polarization is vertical by noting the current flow; the vertical components from the currents in the two upper sides of the triangle add, while the horizontal components cancel.

The objective of the loading is to "lift" the current nodes higher in the vertical space available for the antenna and to make the vertically radiating sides of the antenna more vertical. Both actions will increase

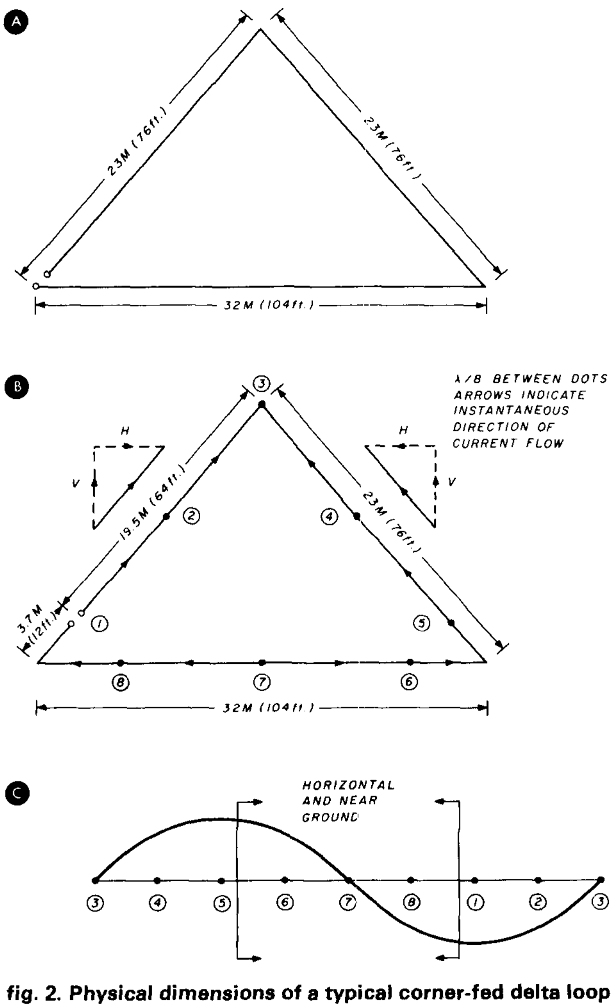


fig. 2. Physical dimensions of a typical corner-fed delta loop antenna (A). True vertical polarization occurs when the feedpoint is one-quarter wavelength from the apex (B). Sketch C shows current distribution.

the low-angle gain for vertically polarized signals. The derivation of the TLDL from a conventional delta loop is shown in **fig. 3**.

The feedpoint resistance for both a conventional delta loop and a TLDL has been measured at W1DTV to be 130 ohms. From this information and from the geometry of the two antennas (and if one assumes sinusoidal current distribution), the TLDL has a gain of 2.3 dB over a conventional delta loop. The dimensions of **fig. 1** have been assumed for this calculation. See reference 3 and **fig. 4** for an explanation of the methods used to arrive at this result.

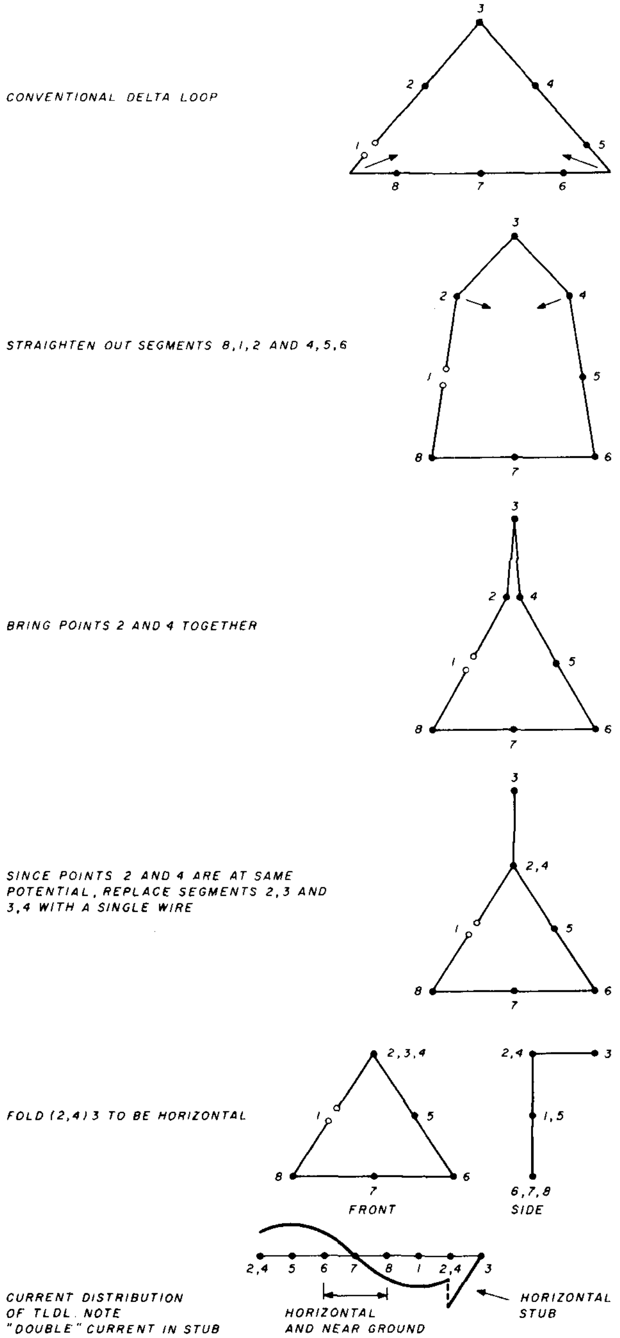


fig. 3. How the top-loaded delta loop (TLDL) is derived from the conventional delta loop. The top loading makes the current nodes higher with respect to ground and increases the vertical polarization from the antenna sides. This improves low-angle radiation.

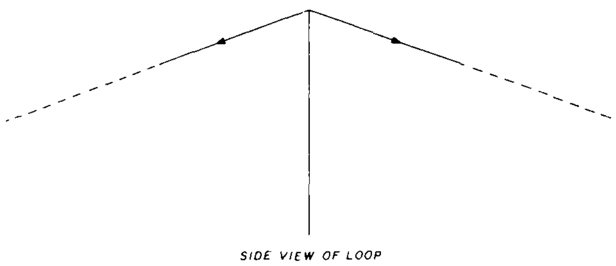


fig. 4. Method of adding the stubs on both sides of the delta loop to reduce horizontally polarized radiation.

The TLDL is truly vertically polarized in a direction perpendicular to the plane of the loop. It is mostly vertically polarized in other directions and exhibits an almost omnidirectional pattern.

loading stub

The loading (or matching) stub is shown in **fig. 3** to be horizontal, but this is rarely possible. At W1DTV it runs to the farthest point on the property and makes about a 60-degree angle with the plane of the loop. The stub should be $\lambda/8$ or 9.8 meters (32 feet) at 3.825 MHz. However, it was necessary to lengthen it to 13 meters (43 feet) for resonance at that frequency. The probable reason for this is that the stub is severely folded back toward the loop; the consequent detuning is overcome by lengthening the stub. This effect is observed in inverted V antennas, where the length must be made longer than would be necessary for a straight dipole.

The stub could be added on both sides of the loop as shown in **fig. 4**. This would virtually eliminate the effect of the stub on the radiated pattern. This method hasn't been tried, and the practical effects are unknown.

The stub can be shortened considerably by means of a loading coil installed in series with the stub at the point where the stub is connected to the triangle apex. See **fig. 5**. The 13-meter (43 foot) stub was reduced to 4.9 meters (16 feet) by the use of a 32-μH loading coil. The loading coil reduces radiation from the stub, but it results in a reduction in antenna bandwidth. The loading coil is a B&W 3029/3905-1,* which is 63.5 mm (2 1/2 inches) diameter by 254 mm

*Barker and Williamson, Inc., Canal Street and Beaver Dam Road, Bristol, Pennsylvania 19007.

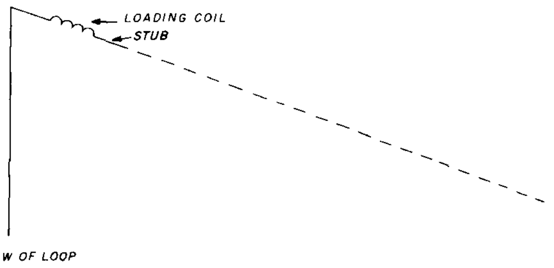


fig. 5. Using a loading coil to reduce stub radiation.

(10 inches) long (6 turns per 25 mm). This coil with the 4.9-meter (16-foot) stub allows the TLDL to resonate anywhere in the 80-meter band by changing the tap position.

matching methods

A common method for feeding delta-loop antennas is to use a quarter wavelength of 75-ohm transmission line between a 50-ohm transmission line and the feedpoint. For a feedpoint resistance of 130 ohms, the vswr at resonance would be

$$\frac{130}{75^2/50} = 1.16{:}1 \qquad (1)$$

which is quite acceptable. Since the conventional delta loop and the TLDL are essentially balanced antennas, it's desirable to use a 1:1 balun at the antenna to prevent antenna currents on the coax feedline.

At W1DTV, a transformer (shown in **fig. 6** and the photo) accomplishes both impedance matching and the unbalanced-to-balanced transformation; it handles the legal power limit quite satisfactorily. The transformer has been evaluated only on 80 meters, but the design could be trimmed to work over several bands. See reference 4 for details on optimizing such designs.

voltage standing-wave ratio

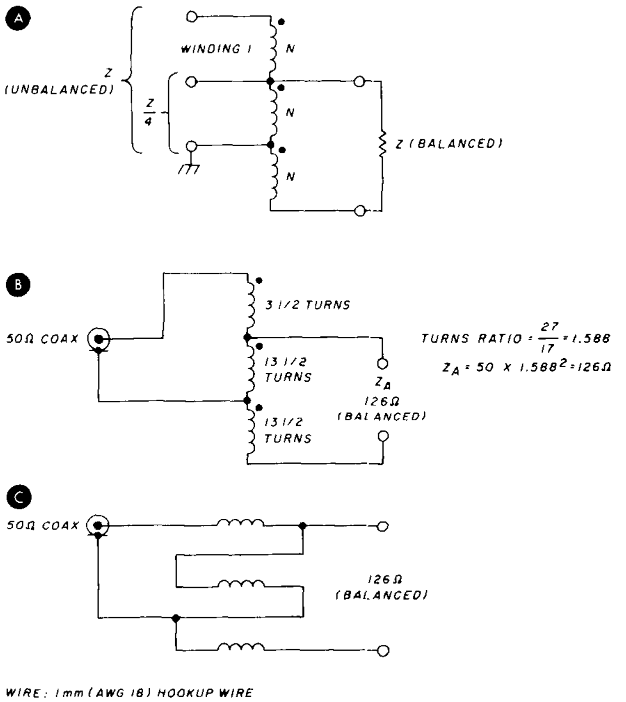
The vswr using the transformer of **fig. 6** and 50-ohm coax is shown in **fig. 7** for the conventional delta loop, the TLDL using a wire stub only, and the TLDL with a wire stub and loading coil. Note that an excellent midband match is obtained for all three cases, but the bandwidth depends on the configuration. The bandwidth of the worst TLDL case (using the loading coil) is substantially better (6.6 times wider) than that of a loaded 20-meter (66-foot), 80-meter sloping dipole,[5] which would require about the same mounting height. The vswr plot of the latter is also shown in **fig. 7** for comparison.

Note from **fig. 7** that the vswr of the TLDL with a loading coil is not unity. The reason is that the inductive loading modifies the current distribution at the top of the antenna, and the feedpoint resistance is changed. For the purist, a more optimum transformer for this case would be one where the 3-1/2-turn winding of **fig. 6** is reduced to 2-1/2 turns. In all cases, the installation involved a steel tower with the antennas supported 1.2 meters (4 feet) from the tower on a boom at the 19.8-meter (65-foot) level. All guy wires were broken with insulators to avoid resonance, and no guying was used above the 10.7-meter (35-foot) level.

concluding remarks

The TLDL antenna performs as well as other similar antennas requiring higher points of support. The design is based on the positioning of the high-current parts of the antenna so that they will provide a primarily vertically polarized radiated signal. The TLDL has substantially more bandwidth than its nearest low-height competitor, the $\lambda/4$ sloping dipole (loaded). Calculations indicate that the antenna should be a good performer, and on-the-air experience has substantiated these results.

A point of caution — if you try this antenna, or any new antenna, take steps to convince yourself that



WIRE: 1mm (AWG 18) HOOKUP WIRE

CORE: AMIDON FERRITE ROD, 12mm DIAMETER, 100mm LONG, μ = 125 (AMIDON ASSOCIATES, 12033 OTSEGO STREET, NO. HOLLYWOOD, CA, 91607)

CONSTRUCTION: 13 1/2 TRIFILAR TURNS WITH 10 TURNS REMOVED FROM ONE WINDING. WRAP WITH VINYL ELECTRICAL TAPE AND TAPE TO THE FEEDPOINT INSULATOR. NO OTHER PROTECTION NECESSARY.

fig. 6. Construction of the matching transformer for the top-loaded delta loop antenna. Sketch A shows the basic principle. Any impedance ratio between 1:1 and 1:4 may be obtained by tapping winding 1. The basic configuration is shown in B. Sketch C shows the schematic and winding logic.

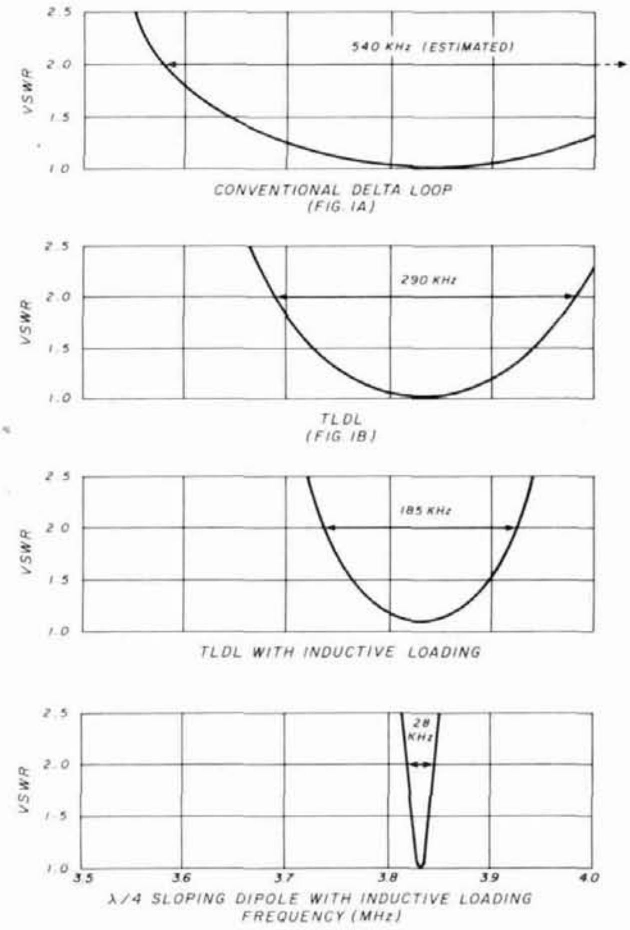


fig. 7. Voltage standing wave ratio and bandwidth of the conventional delta loop, the TLDL, the TLDL with inductive loading, and a quarter-wavelength sloping dipole antenna with inductive loading. Resonant frequency is 3.825 MHz in this model.

other nearby antennas are not significantly influencing its behavior. A considerable amount of interaction between a TLDL, an inverted V, and a sloping dipole, all supported by the same tower, has been observed. The data in this article were taken with the inverted V and sloping dipole removed from the tower.

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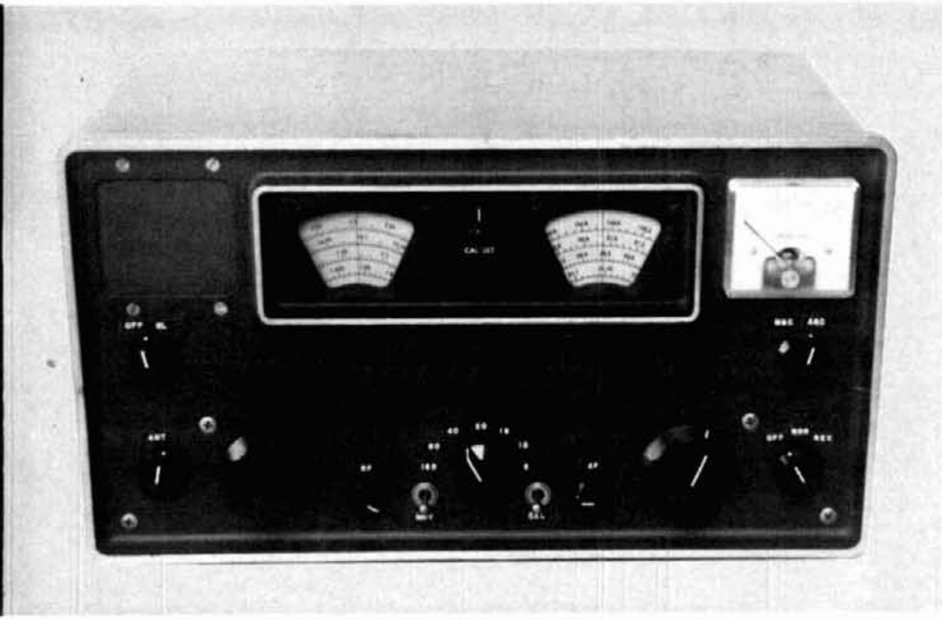
3. F. Witt, "Simplified Antenna Gain Calculations," *ham radio*, May 1978, page 78.

4. J. Sevick, "Broadband Matching Transformers Can Handle Many Kilowatts," *Electronics*, November 25, 1976.

5. G. Hall, "Off-Center Loaded Dipole Antennas," *QST*, September 1974, pages 28-34, 58. (The antenna shown in fig. 5 of the referenced article is the one referred to as the λ/4 dipole. It's a popular radiator for sloping-dipole installations.)

ham radio



updating the vacuum-tube receiver

Rejuvenate that old vacuum-tube receiver by replacing the tubes with equivalent transistor stages

All of the communications receivers made in the 1940s and 1950s, and most of those made in the 1960s, were vacuum-tube designs. Receivers manufactured during those decades were intended for the amateur modes then popular — a-m and CW — and they usually performed quite well on a-m. Inclusion of the beat frequency oscillator made possible reception of ssb and CW, but, by today's standards, much was left to be desired. These older receivers can often be picked up at flea markets or swap meets for a small fraction of their original cost and, properly modernized, can be made to equal or even surpass the performance of new models selling at many times the price.

Probably the two worst shortcomings of older tube-type receivers are their lack of selectivity and their warm-up drift. The drift results mainly from temperature changes brought about by the large amount of heat generated by the tubes. Although much can be done with tube receivers to minimize temperature drift,[1] the best solution is to abandon tubes altogether and substitute transistors. Modern field-effect transistors have characteristics so similar to those of tubes that often a direct one-for-one replacement is possible with only minor circuit and supply voltage changes.

Any of several possible routes can be taken when transistorizing a tube receiver. Probably the most conservative (some would say cowardly) approach is to proceed one stage at a time; that is, substitute a transistor circuit for a vacuum tube stage and perfect its performance before moving on to the next stage. This strategy makes debugging easy, but, because of different impedance and signal levels, tubes and transistors do not always interface easily. Sometimes

By Fred Brown, W6HPH, 1169 Los Corderos, Lake San Marcos, California 92069

a transistor stage will have to be modified when a following or preceding stage is changed over from tube to transistor.

Next to frequency drift, the greatest shortcoming of these old a-m receivers is usually their poor skirt selectivity. The old single-pole crystal filters and *Q*-multipliers gave good selectivity at about 6 dB down, but the skirts were very wide at 30 to 60 dB down. The rectangular passband so necessary for today's ssb communications calls for a mechanical or multi-pole crystal filter.

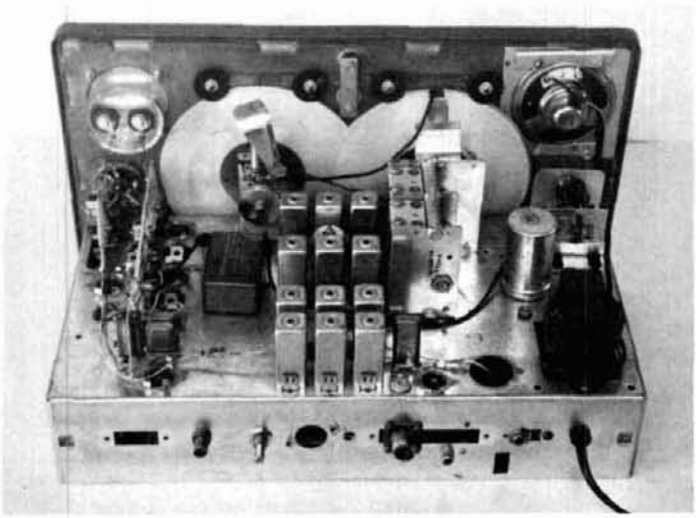
If your receiver is a single-conversion general-coverage type, you will probably have to stick with the original intermediate frequency. This means a mechanical filter if the i-f is below 1 MHz. A ham-band-only receiver, however, can be converted to practically any intermediate frequency so long as the i-f does not lie inside or near an amateur band. (It's very hard to make a superhet that will tune through its own intermediate frequency.)

Generally speaking, a high intermediate frequency is to be preferred, because it improves image rejection. Furthermore, if the local oscillator operates on the low side of the signal, a high i-f means a lower LO frequency, and this contributes to frequency stability. Today, the availability of high-frequency multi-pole crystal filters makes possible single conversion designs with high intermediate frequencies, thereby avoiding the spurious response, birdie, and cross-mod problems of dual-conversion designs.

modernizing the HQ110A

To start this project, I tore out all the old circuits and started with a clean slate, leaving intact only the front-end tuned circuits and bandswitch. Most of the tube sockets were left, as they make handy tie-points for the new wiring. There is something to be said for building from the ground up, but most of the work with sheet-metal and the mechanical drudgery is avoided if you rebuild a commercial receiver. Also, it is hard for the average amateur to duplicate the professional appearance and calibrated dials of a manufactured receiver.

Rear view of the HQ110A. The i-f amplifier, detector, first audio, and S-meter circuits are built on the circuit board that runs along the left side of the chassis.



The HQ110A is a good example of how an intermediate frequency can be radically changed. Originally, the receiver was dual conversion on all bands above 80 meters, with a first i-f of 3035 kHz and a second i-f of 455 kHz. I happened to have a McCoy Golden Guardian 9-MHz ssb filter. I wanted to move the i-f to this frequency and make the receiver single-conversion on all bands. This took some doing; it meant all the original i-f transformers had to be scrapped, and all the local oscillator coils rewound. The result was worth it. The transistorized receiver has much better selectivity, frequency stability, and image rejection than the original. And the existing dial calibration holds on every band, even better than in the tube version.

Power supply and audio. A good place to start your conversion to solid state is in the power supply; you won't be able to try out any transistor circuits without low-voltage dc. Although the old power transformer will be less than ideal for a transistor power supply, it is possible to obtain low voltages from the 5.0- and 6.3- volt filament windings. For instance, a full-wave rectifier on the 6.3-volt winding will provide 9 volts dc, and a full-wave doubler about 18 volts dc. In this receiver, however, I replaced the old power transformer with a more appropriate one having a 16-volt center-tapped secondary.

Both positive and negative supplies were needed, since dual-gate MOSFETs, like vacuum tubes, normally require a negative AGC voltage. Although it is possible to use FETs with positive-only voltages, the AGC system is simplified if a negative supply is available.

The positive supply (see **fig. 1**) is filtered by a conventional L-C filter which uses an 88-mH toroid for a filter choke. The resulting unregulated 19 volts is used for the audio output stage and diode bias. A conventional emitter-follower regulator, Q1 and CR5, is used to regulate the positive supply to +12 volts dc for the remainder of the receiver.

Most designers would use a class B integrated circuit for the audio output, but I prefer the class A arrangement shown in **fig. 1**. This circuit has proven reliable in many different applications. It has excellent dc stability due to the negative-feedback biasing arrangement, and low distortion because of the negative-audio feedback around the output transformer.

Detector, i-f, and AGC systems. The 9-MHz i-f signal from the McCoy filter is applied to the first i-f stage, Q5. This common emitter stage is R-C coupled to the dual-gate MOSFET second i-f stage. R-C coupling is used for simplicity and to avoid the need for neutralization. The added gain that could be obtained from transformer coupling was not needed.

The i-f transformers, T3, T4, and T5, are ordinary fm transistor radio transformers which are shunted with sufficient capacity to resonate at 9 rather than 10.7 MHz. The different resonating capacitors are due to the use of different brands of transformers.

The third i-f stage is an emitter-coupled pair, Q7 and Q8. The emitter of Q7 also drives the AGC amplifier, Q9 and Q10. It was not possible to derive the AGC signal from the collector circuit of Q8 because the BFO signal is too strong at this point and would swamp the AGC. An audio-derived AGC system would have been easier, but in this receiver I wanted AGC that would work on a-m carriers.

The AGC signal is rectified by CR9, with the resulting dc applied to Q11. An emitter follower is used because its low output impedance allows C1 to quickly charge through CR10. In any single-sideband AGC system it is important to have a short attack time to avoid "popping" at the beginning of a word. One inherent limitation on attack time is the delay caused by the crystal filter. Envelope delay time in a bandpass filter is determined by the skirt steepness ratio: the steeper the skirts, the longer the delay. The McCoy filter, having exceptionally steep skirts, has an inordinately long delay time, and for this reason it was not possible to use as much AGC feedback around the filter as would have been desirable. The AGC control voltage to the r-f and mixer stages (AGC 2) is attenuated by the voltage divider, R4-R5. In addition, AGC 2 also controls the S-meter driver through the S9 adjustment pot R4.

The product detector, CR7-CR8, requires push-pull BFO drive which is provided by the center-tapped secondary of T6. The BFO is switched by means of diode switches for upper or lower sideband.

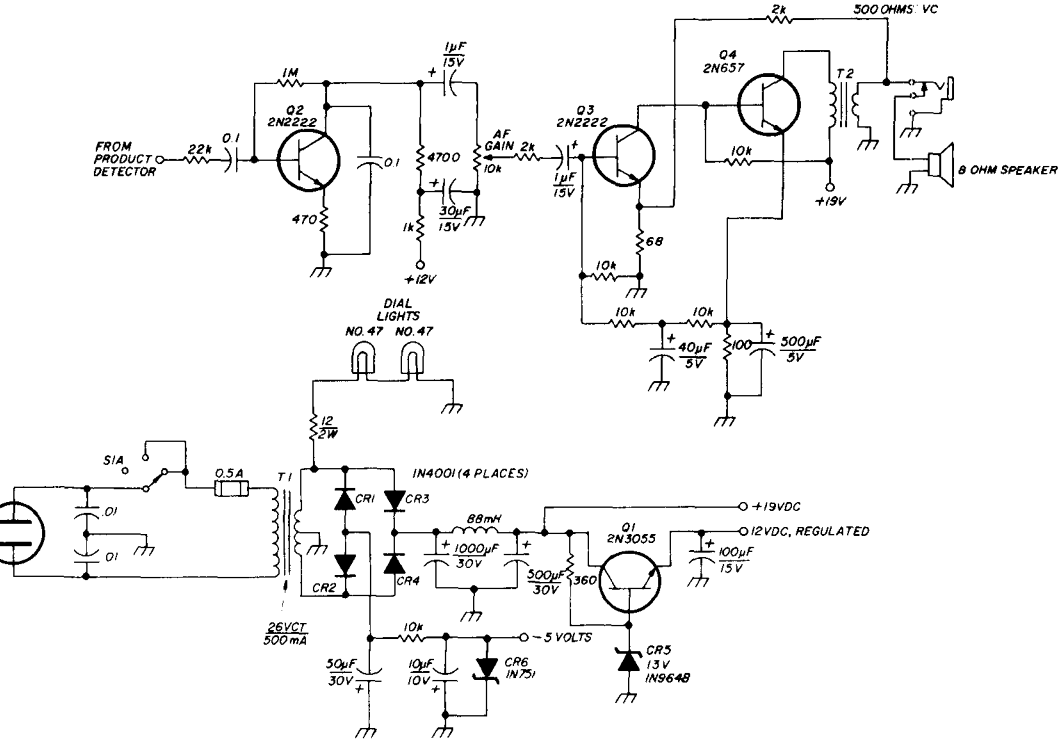


fig. 1. Schematic diagram of the revised power supply and audio circuit. Q1 and Q4 should be mounted on heatsinks. The audio stage will provide 100 mW of undistorted audio output. In addition, the use of Class-A transistors *vs.* an IC may prevent destruction of the devices if the output is shorted.

Local oscillator. In the interest of best frequency stability, the local oscillator operates on the low frequency side of the signal on the 20-, 15-, 10-, and 6-meter bands. On 160, 80, and 40 meters, the oscillator is 9 MHz higher than the signal frequency. This arrangement automatically takes care of sideband switching; upper sideband is received on 20 meters

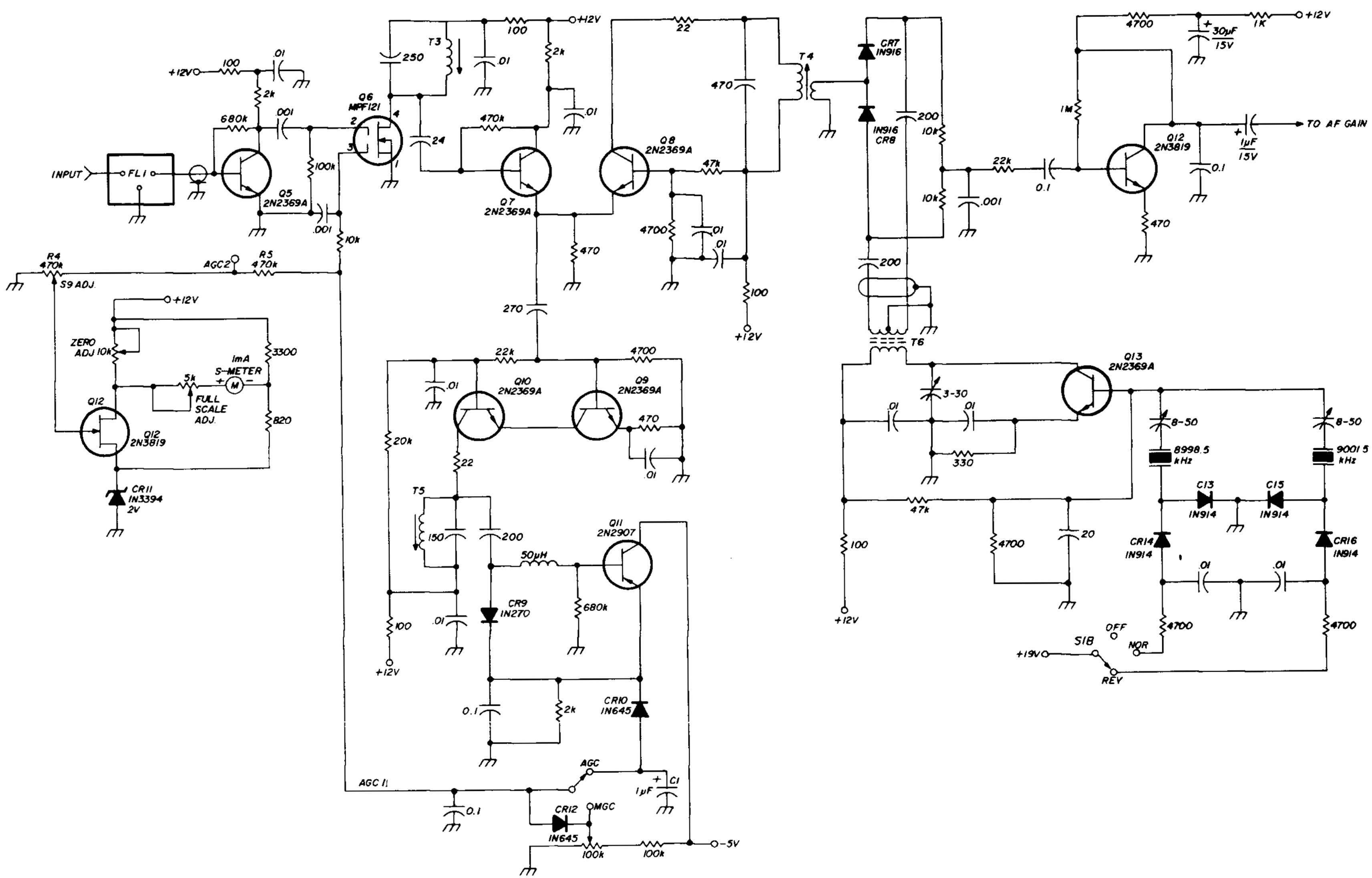
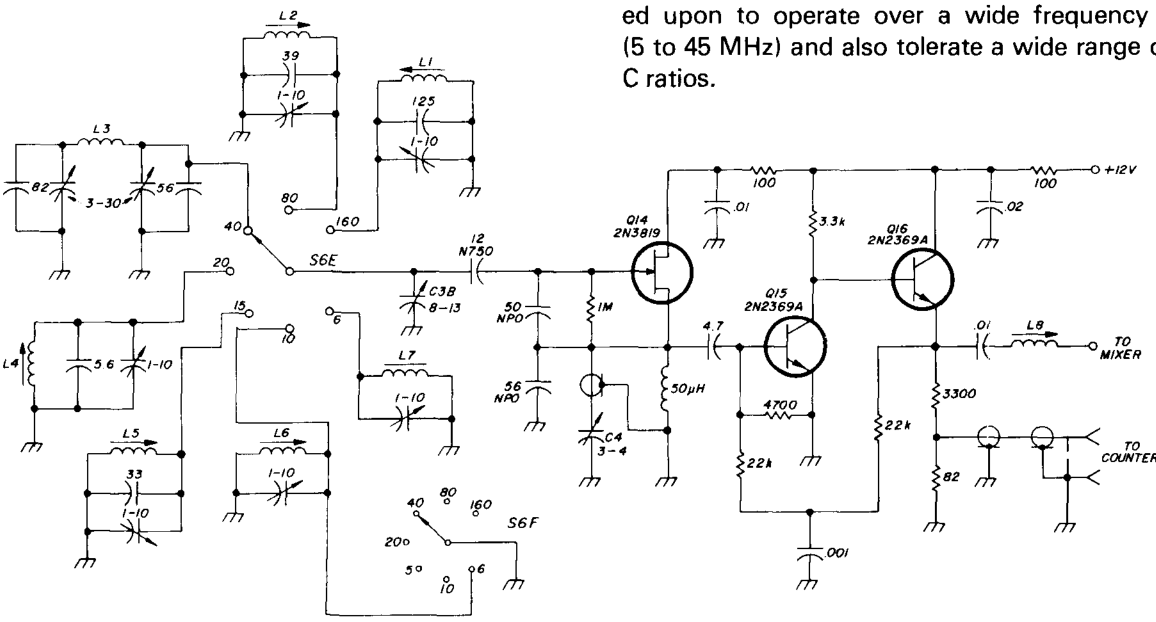


fig. 2. Diagram of the i-f amplifier, detector, BFO, AGC, and S-meter circuits. FL1 is a McCoy Golden Guardian 9-MHz ssb filter. T3, T4, and T5 are 10.7 MHz i-f transformers from fm-style radios. In the case of T3 and T5, where only one winding is shown, the primary and secondary have been connected in series so as to provide maximum inductance. T6 is wound on a T-50-2 core with 33 turns of no. 30 (0.25mm) AWG on the primary and 8 turns of no. 30 (0.25mm) AWG on each side of the center tap for the secondary. Normally, R4 is used to set the S-meter for an S9 indication with 100 μV applied to the receiver.

and the higher bands and lower sideband on 40 meters and lower, as is the accepted standard. The BFO switch can therefore be labeled NORMAL and REVERSE without any reference to the bandswitch position.

Table 1 gives the required oscillator tuning range of each amateur band. On three bands — 20, 15, and 10 meters — the HQ110A dial calibration is wider than the amateur band limits. **Table 1** also gives the *LO tuning range* as a percentage of the lowest oscillator frequency on that band. Notice that the percentage tuning range varies from a maximum of 10.53 per cent on 10 meters to a minimum of 1.85 per cent on 160 meters. The oscillator percentage tuning range is determined by the amount of fixed capacity which shunts the tuning capacitor; the smaller this capacity, the greater the range.

To make the dial calibration come out right, both the inductance and shunt capacitance of the LO tank must be correctly chosen. *In the HQ110A, the oscillator* tuning capacitor (only one of two sections is used) has a minimum capacitance of 8 pF and a maximum of 13 pF, or a range of only 5 pF. This capacitor must be shunted with a total of about 134 pF to provide the 1.85 per cent tuning range for the 160-meter band (10.8 to 11.0 MHz). On ten meters, however, the tuning range expands to 10.53 per cent, and the total shunting capacitance must be reduced to a mere 23 pF. This 23 pF includes the 8 pF minimum of the tuning capacitor, and a few pF of stray and coil-distributed capacitance, and a few pF for the trimmer. When all this is added up, the shunting capacitance left for the remainder of the circuit will have to be limited to a maximum of about 12 pF in order not to exceed the 23-pF total. This is one restriction on the oscillator circuit. In addition, the oscillator is called upon to operate over a wide frequency range (5 to 45 MHz) and also tolerate a wide range of L to C ratios.



L1 14 turns no. 24 (0.5mm)	L4 60 turns no. 32 (0.2mm)	L7 7.5 turns no. 24 (0.5mm);
L2 17 turns no. 28 (0.3mm)	L5 17 turns no. 28 (0.3mm)	L8 17 turns no. 24 (0.5mm);
L3 20 turns no. 22 (0.6mm);	L6 16 turns no. 28 (0.3mm)	

fig. 3. In the local oscillator, C3B is one section of the main tuning capacitor. Inductors L1, L2, L4, L5, and L6 are wound on the original 6.5 mm (¼") slug-tuned form; L3 is wound on a T-68-6 toroid; L7 and L8 are wound on 6.5 mm (¼") slug-tuned forms. Most of the components can be mounted on a small circuit board.

Many different oscillator circuits were tried in search for one that would meet these stringent requirements and would also be stable in frequency. The circuit shown in **fig. 3** worked best. It is an fet version of the Seiler oscillator,[2] followed by a two-stage buffer amplifier. **Table 1** shows the excellent frequency stability of this oscillator for supply voltage changes of one volt (from 11.0 to 12.0 volts). Since the 12-volt supply is fairly well regulated, it's possible to vary the line voltage from 80 to 130 volts ac with no noticeable change in beat note. Try to do that *with your tube receiver!*

Theoretically, this circuit will oscillate with *any* L to C ratio in the tuned circuit. The only limitation on L-to-C ratio is Q; the lower the ratio, the higher the tank circuit Q must be to sustain oscillation. The lowest L-to-C ratios occur on those bands with the smallest percentage tuning range, 160 and 40 meters. On 40 meters, it was not possible to get enough

Q with the original 6.5-mm (0.25-inch) slug-tuned coil form. Therefore, a toroidal coil was used for this band. Since toroids are not adjustable, the circuit arrangement shown in the schematic was adopted. The capacitor in series with the coil (C2) effectively trims the inductance.

On 160 meters, it was possible to wind the coil on the original coil form and still get enough Q, but just barely. All the other coils, except 6 meters, are wound on the original forms and mounted in the original shield cans.

Notice in **fig. 3** that the 10-meter coil is shorted by S6F when the bandswitch is in the 6-meter position. This is because the ten-meter coil happened to resonate near 42 MHz, which, unfortunately, was within the oscillator tuning range on the 6-meter band and caused a back "suck-out" as the oscillator was tuned past this resonant frequency. Also, unfortunately, there was no unused contact on the bandswitch in the 6-meter position, so one had to be added to do this shorting job. This, however, was the only modification that was necessary to the original bandswitch.

The untuned buffer amplifier is made up of a common-emitter stage (Q15), for amplification, directly coupled to an emitter follower (Q16) for low output impedance. As with any amplifier, there is an upper-frequency limit where gain begins to fall off. In this case, output is down about 3 dB at 19 MHz (10 meters), but on 41 MHz (6 meters), it is down about 12 dB. For this reason, a compensating coil (L8) is placed in series with the output. This inductance forms an L network in conjunction with the mixer input capacitance, thereby boosting the 41-MHz LO voltage at the mixer gate. On the lower bands it has little effect.

The 82-ohm resistor in series with the emitter resistor of Q16 delivers a local oscillator signal to a phone jack located on the rear apron of the chassis. By plugging a frequency counter into this jack, you can have a poor-man's digital readout. Of course, it is necessary to add or subtract 9 MHz from the counter reading, but in practice it is easier to simply ignore the digits to the left of the decimal point and mentally substitute the appropriate digits for the band being received. For example, the receiver is on forty meters and the counter reads 16.238. Obviously, this means 7.238 MHz. Remember also, that the counter reading corresponds to the center of the passband, which is typically 1500 Hz higher or lower than the carrier frequency, depending on which sideband is being received.

The tuning rate of the HQ110A was a bit fast for ssb, especially on 10 and 6 meters. For this reason, a clarifier, or ultra-fine tuning control (C4 in **fig. 3**) was added to the LO circuit. This variable capacitor is mounted in the hole formerly occupied by the function switch and is turned by the large, function-switch knob, which is ideal for the purpose.

A very small capacitance range is needed for the clarifier capacitor. It could have been made from a small variable by removing all but two plates, but in my case it was fabricated from the remains of an old wire-wound pot. The resistance element and wiper arm were removed and a semi-circular plate was soldered to the rotor shaft. Another semi-circular plate was fastened to the Bakelite case to form the stator. The resulting capacitor has a range of only 1 pF and produced the frequency ranges given in the fifth column of **Table 1**.

table 1. Frequency ranges for the new local oscillator in the rebuilt HQ110A.

band MHz	oscillator range MHz	oscillator percentage range	oscillator stability Hz/volt	clarifier range Hz
1.8 - 2.0	10.8 - 11.0	1.852	169	420
3.5 - 4.0	12.5 - 13.0	4.0	225	870
7.0 - 7.3	16.0 - 16.3	1.875	165	510
14.0 - 14.4	5.0 - 5.4	8.0	340	1100
21.0 - 21.6	12.0 - 12.6	5.0	257	1000
28.0 - 30.0	19.0 - 21.0	10.53	400	3200
50.0 - 54.0	41.0 - 45.0	9.76	280	7300

rf and mixer stages. Because dual-gate MOSFETs make such ideal replacements for pentode vacuum tubes, no radical changes were necessary in the receiver rf and mixer stages shown in **fig. 4.** None of the antenna or rf stage coils were changed, with the exception of the tap on the six-meter antenna coil: it was moved up to three turns above ground. Both rf and mixer stages, Q17 and Q18, are RCA40673 dual-gate MOSFETs. For simplicity, coils for only the 6-, 10-, 15-, and 20-meter bands are shown in **fig. 4**. Experimentally, it was found that receiver sensitivity on these four bands could be optimized for a 50-ohm antenna by placing a proper size capacitor in series with the antenna input.

When the receiver is tuned to 28.0 MHz, the image frequency happens to be 10.0 MHz, a fact that is exploited with a minimal increase in complexity to provide 10-MHz reception of WWV. For WWV, S5 in **fig. 4** shunts the 10-meter rf stage coil with trimmer C5. This capacitor resonates the 28-MHz coil down to 10 MHz, with the WWV signal from the antenna coupled into the mixer through the series-tuned circuit, L9-C6. The rf stage is not used in this mode; WWV is normally so strong that an rf stage is not really needed.

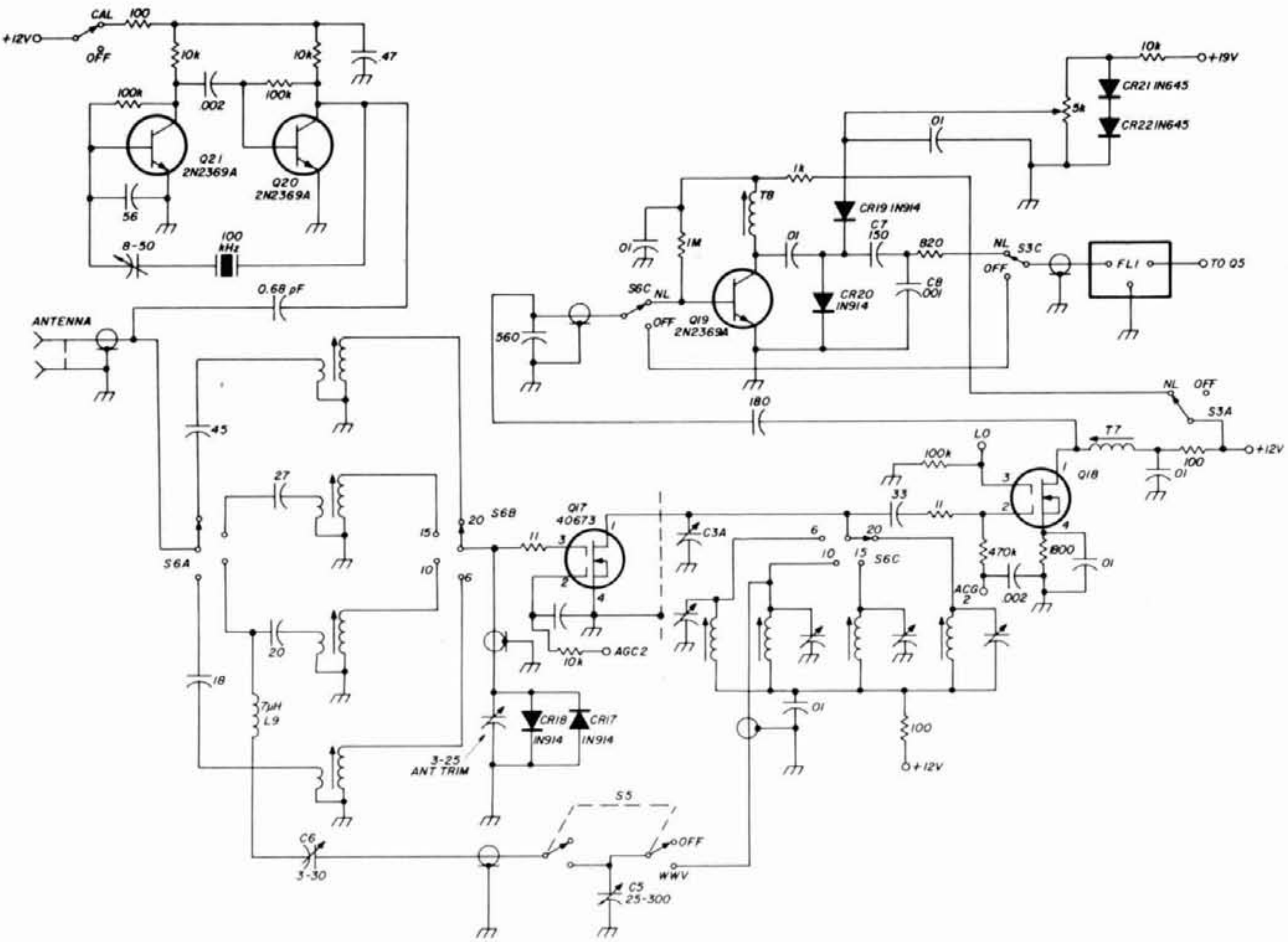


fig. 4. Schematic diagram of the rf, mixer, noise limiter, and crystal calibrator circuits. The antenna and rf stage coils and trimmers are unchanged from the original receiver. For simplicity, only four bands are shown. L9 is 7 μH, made by close winding 30 turns of no. 24 (0.5mm) AWG wire on a 4-mm (5/32-inch) diameter ferrite rod.

Noise limiter and crystal calibrator. I wanted some kind of first-line defense against impulse noise without resorting to all the complexity of a noise blanker. Thus, the simple noise limiter in **fig. 4** was included. In order to be effective, the limiter should clip the noise pulses before they become stretched out by the crystal filter.

The limiter was placed between the mixer and sideband filter and has the option of being switched in or out. The common-emitter stage, Q19, amplifies the noise pulses up to a level where they can be clipped by the shunt limiter, CR19-CR20. These silicon diodes are biased by R6 to a point where they are just beginning to conduct. After being clipped, the signal is brought back down to its original level by means of the capacitive voltage divider, C7-C8.

I make no claim that this limiter can compete with a good noise blanker, but it can make the difference between copy and no-copy. Only high-amplitude noise pulses are clipped. It takes about 50 microvolts at the antenna before clipping begins to occur. This means weak noise pulses are unaffected, but the weak pulses are not normally a problem because they are reduced in amplitude after being stretched out by the i-f filter. For best results, the AGC should be turned off when using the noise limiter.

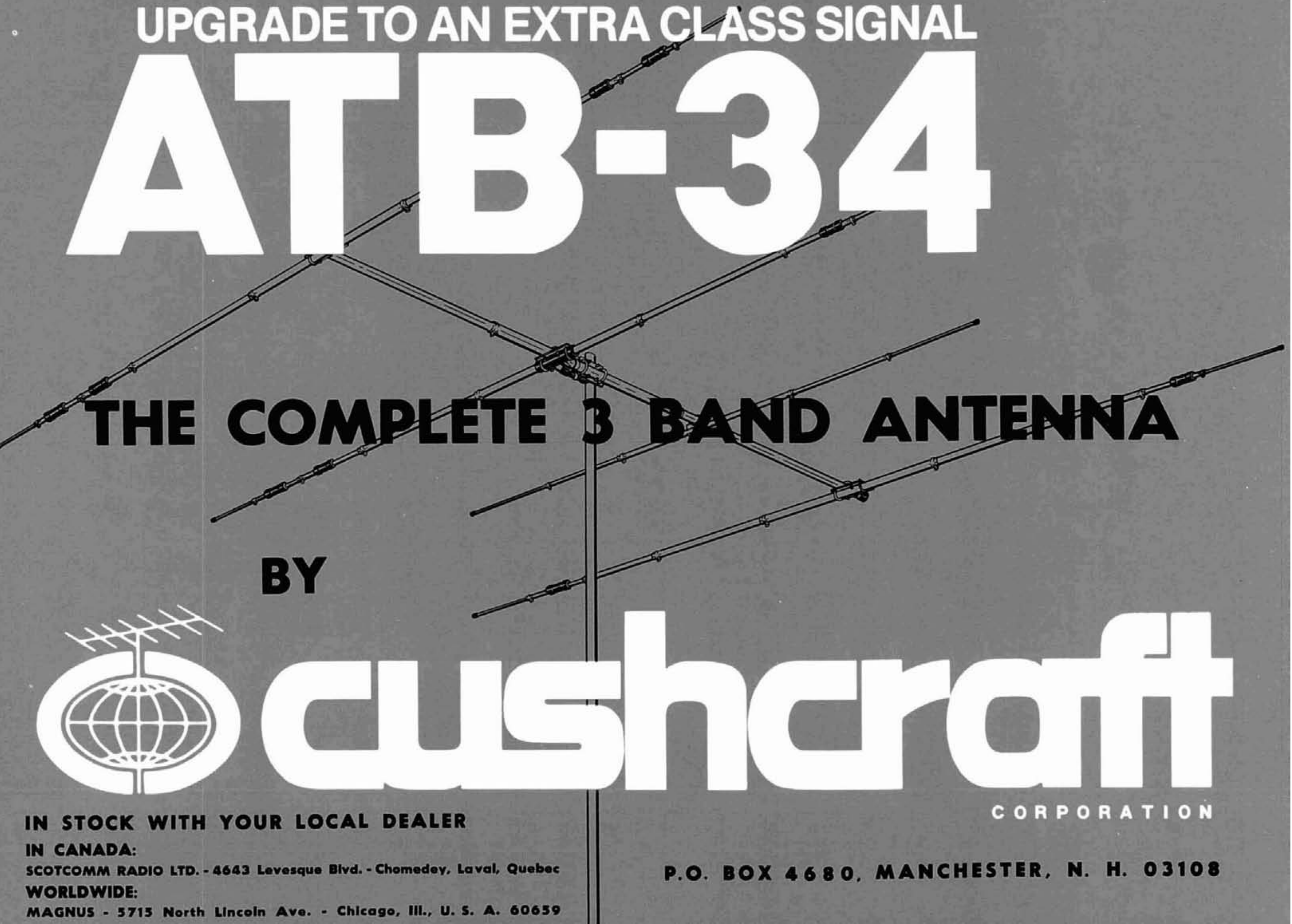
Of course, the greatest shortcoming of this type of limiter is that it is highly susceptible to cross-mod from strong in-band signals. If you have both high noise and strong signals at the same time, you must choose between cross-mod and impulse noise, depending on which is worse.

The 100-kHz crystal calibrator, Q19 and Q20, uses the original 100-kHz crystal and trimmer capacitor. The circuit is patterned after one in the Atlas 210X; it provides plenty of harmonic strength right up through 54 MHz.

references

1. Fred Brown, W6HPH, "Souping Up the Old Receiver," *CQ*, January, 1970, page 15.
2. E. O. Seiler, W8PK, "A Low-C Electron-Coupled Oscillator," *QST*, November, 1941, page 26.

ham radio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double-stub tuner for 1296 MHz

Feedline and matching problems at 1296 MHz are easily overcome by using this simple but effective double-stub tuner

Shortly after the military APX-6 became available on the surplus market in the 1950s, a number of magazine articles were published which described conversion of the unit to the 1215-1300 MHz amateur band. For a time there was considerable interest in this band; here in the San Fernando Valley there were about seven stations active for a year or more. Eventually interest declined, until there were only two of us left — K6BV and myself. We had 85 MHz of amateur frequency spectrum with almost no takers. That's too bad, because this band is ideal for repeaters, wide-band television, satellite communications, and amateur experimentation.

Why is the use of this band so meager? Probably the main reason is that there is no equipment available for the band. Also, it requires the application of techniques quite different from those used on the lower frequencies; of necessity, striplines and cavities must be used. These take some getting used to by newcomers to the amateur bands above 1000 MHz.

Another aspect that has not received too much attention is transmission lines. Losses in coaxial cable are tremendous at this frequency. Standing waves are a disaster if they appear on the line. The best RG-8/U coax has a loss of 8 dB per 30 meters (100 feet) at 1296 MHz. Any reflection at the antenna is power lost in all but the shortest lines, and putting a radiator at the top of a tower does not promote a short feedline. On the low-frequency amateur bands a reasonable reflection of power at the antenna does not represent any appreciable loss in radiated power; if nothing gets hot, and the transmitter loads, the power has to radiate. Reflected power makes a round trip at low frequencies. At 1296 MHz, the reflected power suffers the 8-dB loss on the trip back to the transmitter and again on the second trip to the antenna. The net result is that reflected power is power lost, and power is hard to come by at 1296 MHz.

Measurement of power and standing waves on the feedline was a serious problem at my station for several years. A field-strength meter is easy to make for 1296 MHz, and setting it up in front of the antenna gives a good indication of relative power levels, but it usually turned out that a slotted line was still showing a considerable standing wave on the feedline, even after everything was tweaked for hours on end. Just taking the slotted line out of the line was apt to be a disaster if only because of the change in length of the line. In addition to this problem, the field-strength meter didn't give a true reading of the actual power being radiated. It was strictly a relative measurement, and I could not tell whether the transmitter was running at 5 per cent efficiency or the hoped for 30 to 40 per cent.

Uncertainty over the power output of the transmitter and the unhappy situation with standing waves on the feedline led to several not-so-cheap purchases and also a rather close-tolerance construction project. The first purchase was a Bird model 43 wattmeter with a plug-in covering 1.1 to 1.8 GHz. The second purchase was a Sierra Model 160B-300 dummy load (I couldn't build a suitable dummy load and gave up after a number of attempts). The way to rationalize these purchases is to say that the test gear will not wear out and with much cheaper plug-in elements the Bird 43 wattmeter will work on other amateur bands.

With the wattmeter and the dummy load I could read transmitter power output and pick up reflected

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power when the feedline was moved from the dummy load to the antenna radiator. Things began to fall into place. It became possible to trim the driving dipole for a reasonable match to the line. Single-helix antennas were stubborn, and responded only half-heartedly to matching devices; a twin-helix antenna was much better; a quad helix never seemed to work just right, regardless of the matching scheme I tried.

A nasty development surfaced when the dipole in front of my 2.5-meter (8-foot) dish was moved for field-strength tests and beamwidth measurements. Any change of antenna configuration forced me to retune the driver. The answer, of course, was to find some kind of impedance-matching device that could be put into the line at the antenna and at the transmitter. The double-stub impedance transformer was the logical solution. But, where to find one? The surplus houses I trade with never had a single one that could possibly have worked at 1296 MHz. I had to build one; in fact, I had to make four of them. (Anybody working 1296 is bound to have more than one antenna.)

The impedance-matching range of the double-stub impedance-matching transformer is astonishing. For example, it is no problem at all to bring the feedlines from two dipole antennas together at a T and match the resulting 25-ohm impedance to a 50-ohm line.

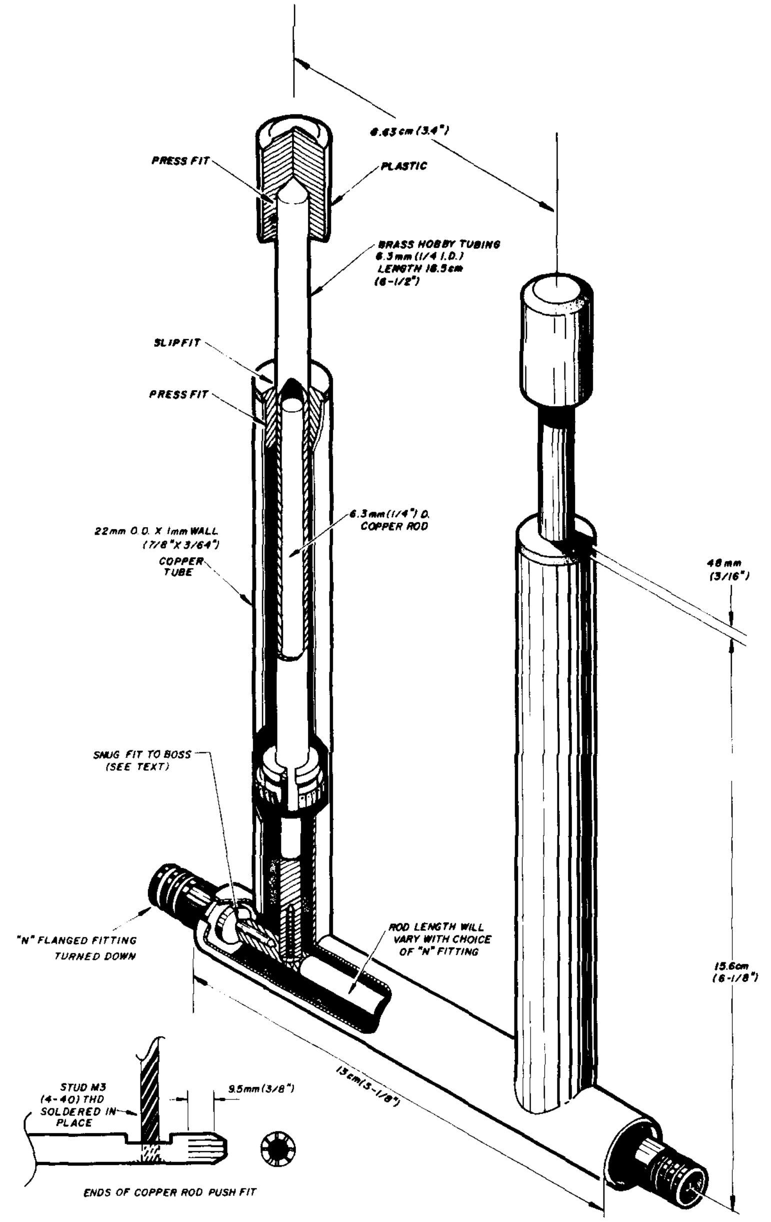

fig. 1. Mechanical diagram of the double-stub tuner. As mentioned in the text, the length of the center conductor of the main line is determined by the type of N connector which is used. Matching is accomplished by the two stubs with sliding shorts which provide adjustable amounts of reactance.

tuner construction

The stubs of a double-stub tuner are normally 3/8 wavelength apart and mounted at right angles to a 50-ohm line section. Insertion loss is not readable on a 0- to 25-watt rf power meter running at half scale. There is just one problem with the tuners: they are hard to make because mechanical tolerances must be held tightly so that the tuning plungers do not bind when they are adjusted. It is not necessary to have a mill to build a double-stub tuner, although there is one minor place where it produces a more professional job. A lathe should be available.

construction

Referring to the cross-sectional sketch (see **fig. 1**), tubing sections used for the stubs can be sawed to length; remember that they must be long enough to reach halfway around the mainline section. Cut and file one end of each stub tube to fit snugly around the main section line. Scribe the *outside* contour of one of the stub tubes onto the main line, 43 mm (1.7 inches) from the center of the main line; scribe a similar contour 43 mm (1.7 inches) from the center toward the other end. These outlines must be parallel along the main line tube. Remember, these scribed contours are the *outside* of the stub tubes. Cut and file openings in the main line tube to match the inside diameter of the stub sections. Final fit must be made with the file strokes at right angles to the line tube and parallel to the axis of the stub tubes, as they will solder in place. This may sound difficult, but it's not, just a bit tedious.

Assembly of the two stubs to the main line must be done accurately. I found the simplest way was to lay the three parts on a piece of 1.9-cm (3/4-inch) plywood and drive nails on both sides of all three tubes to hold them in place. A square will indicate when the stubs are at right angles to the main line. The plywood will keep all three parts in the same plane. Put something heavy on each of the tubes to keep it from shifting during the soldering operation. The stubs are soldered to the main line section with a substantial bead halfway around the joint. The assembly is then turned over in the jig to complete the second half of the soldering job. If the half-soldered

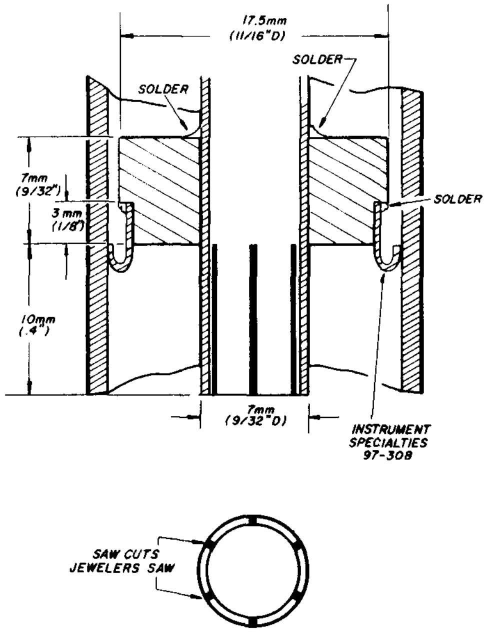

fig. 2. Close-up view of the fingerstock assembly. If brass shim stock is used instead of fingerstock, the size of the brass ring can be changed to compensate for the size difference.

assembly, when flipped over, does not drop into the nail jig without binding, the right-angle alignment of the jig is faulty. If this happens, it is best to start over.

In **fig. 1**, the dimension of the inner conductor of the main line has not been shown. This is because flanged N connectors vary somewhat in design, and the builder must determine this length to match the available fittings. For final assembly, the outer face of both N flanges must be recessed into the main line tube approximately 1.5 mm (1/16 inch). The normally square flanges are turned on a lathe to a diameter allowing a slip fit into the copper tube. The center-conductor's length is set to fit snugly up against the boss at the base of the conductor extension of each N fitting. This is not critical, but recessing of the N connector flanges is necessary to permit a substantial solder bead for maximum strength under torque load from connected equipment.

Once the inner conductor length is established, the ends of the rod are tapered and an 11 mm (7/16 inch) deep hole is drilled in each end. The diameter of the hole is selected to provide a slip fit on the center pin of the connector. With a jeweler's saw, three longitudinal cuts are made on each end of the center-conductor rod, giving six segments at each end. The segments are given a set toward the center to establish a press fit over the center pin of the N conductor (no soldering is possible at final assembly). The longitudinal cuts are 9.5 mm (3/8 inch) deep. If the N conductor pins are more than 6.5 mm (1/4 inch) long, cut them back to 6.5 mm (1/4 inch).

There is one very critical dimension in the inner conductor assembly, the spacing between the points at which the stubs connect. This spacing must match, as closely as possible, the center spacing between the outer tubes already assembled. Measure the distance between the stub centers accurately and transfer this measurement to the center conductor of the main line. (Center-to-center measurement is most easily done by measuring from the left-hand side of one tube to the left-hand side of the other tube, or conversely, from the right-hand side of one tube to the right-hand side of the other tube.) This measurement is transferred to the center conductor of the main line so that the distance from one end of the line to the first stub is equal to the distance from the other end of the line to the second stub.

Clamp the rod in a vise and file flat spots 6.5 mm (1/4 inch) wide over the stub locations. These should be filed one-third of the way through the rod. The faces must be flat and parallel to each other since they affect final alignment of the stubs. Drill and tap each flat for a 4-40 (M3) machine screw, ensuring that the distance between the drilled holes is exactly the same as the distance between the centers of the two stub tubes already assembled. Insert brass 4-40 (M3) machine screws from the bottom so that the threaded end of the screws extends 9.5 mm (3/8 inch) above the face of the flats. The above procedure is for those who don't have access to a mill. (With a mill, sink a 6.5-mm [1/4-inch] end mill 3 mm [1/8 inch] deep, then drill and tap for the screws.)

The center-stub conductors are squared in the lathe, drilled and tapped for the screws already in place in the main line. Screw the stubs onto the 4-40 (M3) studs and check for perpendicular projection from the line and for parallel alignment. If alignment is correct, the screws can be soldered to the main line, after which the excess should be sawed off and smoothed to maintain the shape of the line. The stubs must be removed from the mounting studs for final assembly.

The plungers are made from 6.5-mm (1/4-inch) ID brass hobby tubing, which is a slip fit on the center conductor stubs. The tubing is also split with a jeweler's saw as indicated in **fig. 2**. The segments are pressed inward to give a good sliding fit on the center conductor. The brass ring which supports the fingerstock is bored, turned and soldered in place. The indicated fingerstock is ideal for the job, but brass shim stock may be substituted if the fingerstock is not available. When using shim stock, wrap it around the ring, slit in segments, and solder in place. Give the segments an outward bias and contour the tips to slide rather than dig into the outer conductor.

with the plungers completely in. It is good insurance to wrap the base of the stub tuner lines, where they solder to the main line, with a damp cloth while soldering the N fittings into place.

With these three pieces of gear, directional watt meter dummy load, and stub tuners, 1215 MHz to 1300 MHz becomes one of the easiest of the amateur bands to work on, rather than one of the hardest.

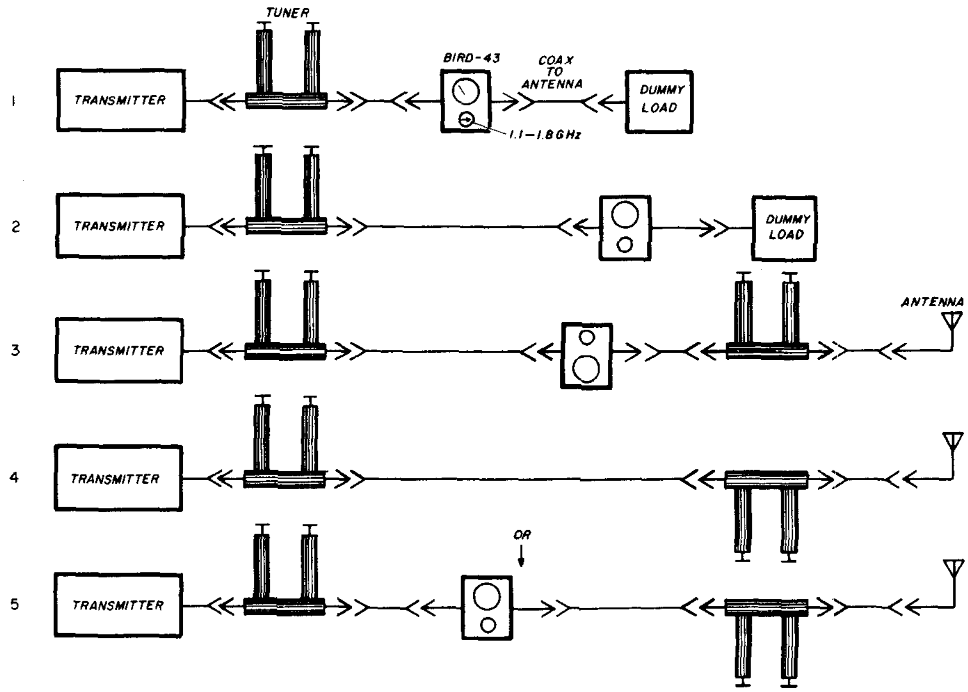

fig. 3. Different configurations for using one or two double-stub tuners.

The plastic fitting at the end of each stub is drilled for a slip fit on the plunger tubing and pressed onto the end of the stub. The plunger knobs may be any type of plastic and are also press fitted onto the plunger tubing.

final assembly

The center conductor of the main line is slipped into the tube, the center conductor stub lines are screwed into place on the 4-40 (M3) screws and set tight with a pair of heavy pliers. The N fittings are pressed into the ends of the main line center conductor and the plungers are slid down over the stub center lines. The plastic guide sleeves are pressed into the open end of the stub tubes and the plungers are run up and down to check for any binding. There should be none, though a slight difference in resistance may be noticed at first as the plungers move full stroke. If no problems show up, the flanges of the N fittings may be soldered with the plungers both fully inserted. This is important, as any misalignment will be less troublesome if the N fittings are soldered

The guesswork disappears in checking cavity or stripline efficiency, antenna match, and line reflections. Comparing one antenna to another becomes routine, rather than the long, drawn-out process of trying to match power into an antenna system in the face of all the line and feed problems that constantly disturb the results.

Different tuner configurations are illustrated in **fig. 3**. As a simple indication of what can be accomplished with one of these tuners, a test set-up was arranged on the bench. Eight watts of power was fed into a dipole that had been trimmed to give a good match to a 50-ohm line when used to drive a 2.5-meter (8-foot) parabolic dish. The wattmeter showed that two watts were being reflected. With a double-stub tuner into the line at the dipole, it took less than ten seconds to wipe out the reflected power by adjusting the stub plungers. This was a 25 per cent increase in radiated power without having to touch the antenna. If aggravation has a price tag, these three pieces of gear are a bargain.

ham radio

TS-520S

THE TS-520S SERIES LITERALLY TOOK THE AMATEUR WORLD BY STORM. NO OTHER RADIO EVER CAUGHT ON SO FAST AND THE REASONS ARE OBVIOUS...EXCELLENT PERFORMANCE CHARACTERISTICS, DEPENDABILITY, FLEXIBILITY, AND A VERY SOLID VALUE FOR THE PRICE. AND NOW THE TS-520S SERIES OFFERS THE MOST COMPLETE LINE OF ACCESSORIES AVAILABLE.

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The TS-520S provides full coverage on all amateur bands from 1.8 to 29.7 MHz. Kenwood gives you 160 meter capability, WWV on 15.000 MHz., and an auxiliary band position. And with the addition of the TV-506 transverter, your TS-520S can cover 160 meters to 6 meters on SSB and CW.

OUTSTANDING RECEIVER SENSITIVITY AND MINIMUM CROSS MODULATION

The TS-520S incorporates a 3SK35 dual gate MOSFET for outstanding cross modulation and spurious response characteristics. The 3SK35 has a low noise figure (3.5 dB typ.) and high gain (18 dB typ.) for excellent sensitivity.

NEW IMPROVED SPEECH PROCESSOR

An audio compression amplifier gives you extra punch in the pile ups and when the going gets rough.

VERNIER TUNING FOR FINAL PLATE CONTROL

A vernier tuning mechanism allows easy and accurate adjustment of the plate control during tune-up.

FINAL AMPLIFIER

The TS-520S is completely solid state except for the driver and the final tubes.
Kenwood has employed two husky S-2001A (equivalent to 6146B) tubes. These rugged, time-proven tubes are known for their long life and superb linearity.

HIGHLY EFFECTIVE NOISE BLANKER

An effective noise blanking cricuit developed by Kenwood that virtually eliminates ignition noise is built into the TS-520S.

RF ATTENUATOR

The TS-520S has a built-in 20 dB attentuator that can be activated by a push button swich conveniently located on the front panel.

PROVISION FOR EXTERNAL RECEIVER

A special jack on the rear panel of the TS-520S provides receiver signals to an external receiver for increased station versitility. A switch on the rear panel determines the signal path . . . the receiver in the TS-820 or any external receiver.

CW-520 — CW FILTER (OPTION)

The CW-520 500-Hz filter can be easilly installed and will provide improved operation on CW.

AMPLIFIED TYPE AGC CIRCUIT

The AGC circuit has three positions (OFF, FAST, SLOW) for optimum operation on CW.

AC POWER SUPPLY

The TS-520S is completely self-contained with a rugged AC power supply built-in. The addition of the DS-1A DC-DC converter (optional) allows for mobile operation of the TS-520S.

EASY PHONE PATCH CONNECTIO

The TS-520S has two convenient RCA phono jacks on the rear panel for PHONE PATCH IN and PHONE PATCH OUT.

The TS-520S retains all of the features of the original TS-520 that made it tops in its class: RIT control • 8-pole crystal filter • Built-in 25 kHz calibrator • Front panel carrier level control • Semi-break-in CW with sidetone • VOX/PTT/MOX • TUNE position for low power tune up • Built-in speaker • Built-in cooling fan • Provisions for four fixed frequency channels • Heater switch.

VFO-520S TV-502S TV-506 SM-220

ting on the 2-meter band.
The **TV-506** is an equally practical way of getting on the 6-meter band, providing 10 watts on SSB and CW.
The **SM-220** is an extremely useful and unique station monitor. It allows you to monitor your transmissions, monitor incoming signals and monitor the amount and strength of band activity* and performs as a general purpose 10 MHz oscilloscope, as well.

*With BS-5 or BS-8 pan display option.

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TS-820S

You can always tell who's running a TS-820S. Its superb quality stands out from all the other rigs on the band ... and when the QRM gets heavy, the TS-820S's adjustable RF speech processor, utilizing a 455-kHz circuit to provide quick-time-constant compression, will get the message through. RF negative feedback is applied from the final to the driver to improve linearity, and third-order products are at least −35 dB. Harmonic spurious emissions are less than −40 dB and other spurs are less than −60 dB. RF input power is 200 W PEP on SSB, 160 W DC on CW, and 100 W DC on FSK. Receiver sensitivity is better than 0.25 μV for 10 dB S/N. The TS-820S is known for its superb receiver selectivity, and its famous IF shift easily eliminates heavy QRM. That's why the TS-820S is the DXer's choice.

See your local Authorized Kenwood Dealer today.

The Perfect Station

SP-820 TS-820S VFO-820 TV-502S TV-506 SM-220

T-599D

The R-599D and T-599D... ...now less than $1,000

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The T-599D transceiver is solid-state with the exception of only three tubes, has built-in power supply and full metering. It operates CW, LSB, USB and AM and, of course, is a perfect match to the R-599D receiver.

The R-599D is the most complete receiver ever offered. It is entirely solid-state, superbly reliable and compact. It covers the full Amateur band, 10 through 160 meters, CW, LSB, USB, AM and FM.

Your station isn't complete if it doesn't include the R-820

R-820

Introducing the ultimate in receiver design . . . the Kenwood R-820.
With more features than ever before available in a ham-band receiver. This triple-conversion (8.33 MHz, 455 kHz, and 50 kHz IFs) receiver, covering all Amateur bands from 160 through 10 meters, as well as several shortwave broadcast bands, features digital as well as analog frequency readouts, notch filter, IF shift, variable bandwidth tuning, sharp IF filters, noise blanker, stepped RF attenuator, 25 kHz calibrator, and many other features, providing more operating conveniences than any other ham-band receiver. The R-820 may be used in conjunction with the Kenwood TS-820 series transceiver, providing full transceive frequency control.

Additional features include: A monitor switch which allows the user to hear his own voice when using associated transmitter. Either VFO control or crystal control on four selectable frequencies. Digital hold . . . locks counter and display while VFO is tuned to another frequency . . . facilitates return to "hold" frequency. RIT/notch control . . . RIT allows receiver to be tuned off frequency, while not affecting transmit frequency when in transceive mode with TS-820S. Notch control tunes notch within IF passband for eliminating interference. Interfering signal remains notched even when IF shift is utilized. Built-in crystal calibrator, settable to WWV, provides signal every 25 kHz. Noise blanker/level control . . . for maximum reduction of noise interference. A transceive/separate switch enables receive VFO to control the receiver and TS-820 (or TS-820S) frequency (or the TS-820 VFO to control both), or, of course, both can function independently.

TL-922A

the most versatile pair on the air

If you have never considered the advantages of operating a receiver/transmitter combination . . . maybe you should. Because of the larger number of controls and dual VFOs the combination offers flexibility impossible to duplicate with a transceiver.

Compare the specs of the R-599D and the T-599D with any other brand. Remember, the R-599D is all solid-state (and includes four filters). Your choice will obviously be the Kenwood.

R-599D

OR 25 WATT OUTPUT

TR-7400A

The fully-synthesized TR-7400A 2-meter FM transceiver operates on 800 channels and features repeater offset over the entire 144-148-MHz range, dual frequency readout, six-digit display, and subaudible tone encoder and decoder. RF output is at least 25 watts!

The TR-7400A 2-meter FM transceiver provides fully synthesized operation, *including 600-kHz repeater offsets,* over the entire 144-148-MHz range. It can operate on any of 800 channels, spaced 5 kHz apart. RF output is at least 25 W, and typically 30 W. A low power position produces 5-15 W (adjustable). Included is a dual frequency readout with large six-digit LED display plus a dial readout. The sub-*audible CTCSS signaling feature may* be used on transmit and receive, or transmit only. Optional tone-burst modules are available. Receiver sensitivity is better than 0.4 μV for 20 dB quieting. Large, high Q, helical resonators minimize interference from outside the band. A two-pole 10.7-MHz monolithic crystal filter provides excellent selectivity. Intermodulation distortion is down more than 66 dB, *spurious rejection is better than* -60 dB, and image rejection is better than -70 dB.

See your local Authorized Kenwood Dealer today, for a demonstration of the fantastic TR-7400A.

TS-600

Experience the excitement of 6 meters. The TS-600 all mode transceiver lets you experience the fun of 6 meter band openings. This 10 watt, solid state rig covers 50.0-54.0 MHz. The VFO tunes the band in 1 MHz segments. It also has provisions for *fixed frequency operation on* NETS or to listen for beacons. State of the art features such as an effective noise blanker and the RIT (Receiver Incremental Tuning) circuit make the TS-600 another Kenwood "Pacesetter".

TRIO-KENWOOD COMMUNICATIONS INC.
1111 WEST WALNUT/COMPTON, CALIFORNIA

1.5 GHz divide-by-four prescaler

Extending the range of your counter is made easy by this simple divide-by-four 1500-MHz prescaler

In the last few years frequency counters have developed from bulky, slow, costly units affordable only by well-heeled laboratories, into small, high-frequency instruments at a price easily within the reach of the average ham. Quite a few manufacturers now offer counters with 1-Hertz resolution to 50 or 60 MHz, low power drain, and a price in the neighborhood of $100. In addition, many articles have appeared in the amateur magazines describing the construction of frequency counters and prescalers; recently, an entire issue (February 1978) of *ham radio* was devoted to the subject of frequency counters.

The upper frequency limit of every counter is set by the maximum operating speed of the input divider stage. This first divider is usually a fast prescaler — a fixed, asynchronous divider operating with the main divider chain. The most popular low-cost prescalers use the 11C90 digital divider, which has a typical maximum operating frequency of about 600 MHz. IC prescalers with higher operating speeds are available, but their high prices have kept them from finding much amateur use.

However it is now possible to inexpensively extend a counter's operating frequency range to well beyond 1 GHz, with a recently introduced ECL (emitter-coupled logic) divide-by-four prescaler. This device is the Motorola MC1697, a very fast ECL prescaler which has a typical operating frequency of 1600 Mhz. Thus, a prescaler using this circuit will extend the range of a 400-MHz counter to above 1.5 GHz. It will operate with signals as low as 1 mW, requires only a single dc supply, and will drive 50-ohm loads. The MC1697 is a low-cost, plastic-packaged device that sells for less than $18.

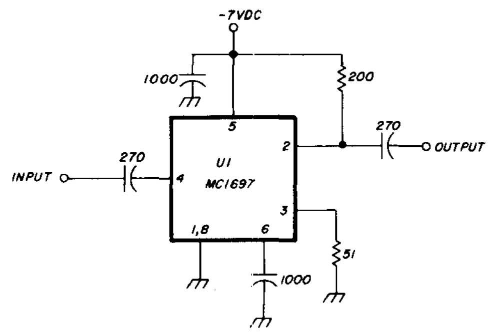

fig. 1. Schematic diagram of the divide-by-four prescaler using the Motorola MC1697. Chip capacitors are preferred for the input and output coupling capacitors, though silver micas may be used if the other types cannot be obtained. If the IC should free-run, the oscillations can be stopped by connecting a 10k resistor across pins 4 and 5.

By Jerry Hinshaw, N6JH, 2500 Medallion #193, Union City, California 94587

circuit description

The schematic diagram of the prescaler's circuit is shown in **fig. 1**. The signal to be counted is fed through C1 to the MC1697, which contains two divide-by-two stages in cascade. The input to the first counter is internally biased to accept zero-mean signals so that no external set-up bias is needed. The second flip-flop has complementary outputs. One of these is fed to the output connector through C2, the other is terminated in a 51-ohm resistor. The 200-ohm resistor terminates the open emitter of the output emitter follower.

A single dc power supply is required to provide −7 volts at approximately 60 mA. A simple power supply for this prescaler module can be built using an adjustable voltage regulator. The design of power supplies using this type of regulator has been discussed by K5VKQ.[1]

construction

The prescaler is built on the small printed circuit board shown in **fig. 2**.* Double-clad, 1.6-mm (1/16-inch) glass-epoxy board is used. One side forms a ground plane, which provides for low-impedance grounding and permits the use of microstrip lines for the input and output.

All components, with the exception of C1 and C2, are mounted on the ground plane side of the board. If chip capacitors are used for C1 and C2, they should be soldered directly across the gaps in the microstrip lines. If silver mica capacitors are to be used, they may be mounted on the ground plane side of the board with the rest of the components. The use of chip capacitors is preferable, as they will provide lower loss and less inductance than capacitors with wire leads.

Be sure to leave enough clearance under the bodies of the two 1000-pF bypass capacitors so that the lead can be soldered to the ground plane at the points indicated by Xs on **fig. 2C**. One end of the 51-ohm resistor is connected to the top side of the board.

If a socket is used to mount the MC1697, it must be the type that can be soldered to both sides of the board so that pins 1 and 8 can be properly grounded. Molex or swage-in pins will work here, or the IC can be soldered in the circuit board. Soldering the IC is the best choice in terms of electrical and thermal performance, but it does make changing the device difficult. If you are going to purchase only one IC (rather than buy several to select the fastest), you may as well solder it directly to the board.

*An etched, drilled, and plated glass-epoxy circuit board is available for $6.00 postpaid, from H&M Microwave, Post Office Box 185, Montrose, California 91020. California residents please add 6 per cent sales tax.

The board was designed to fit into a Bud CU124 cast aluminum box, although it will also fit into a 5.1 × 7.6 × 10.2-cm (2 × 3 × 4-inch) sheet-metal box. The cast box provides better grounding and rf shielding. The board fits into the box as shown in **fig. 3**. The component side is mounted facing down; this is done so that the BNC input and output connectors can be directly soldered to the microstrip lines after the board has been installed in the box.

The board itself is mounted in the box with four 9.5-mm (3/8-inch) standoffs located at the four mounting pads near the corners of the board. The standoffs should be soldered to the ground plane and then securely clamped to the box with screws.

If an external power supply is used, the lead may

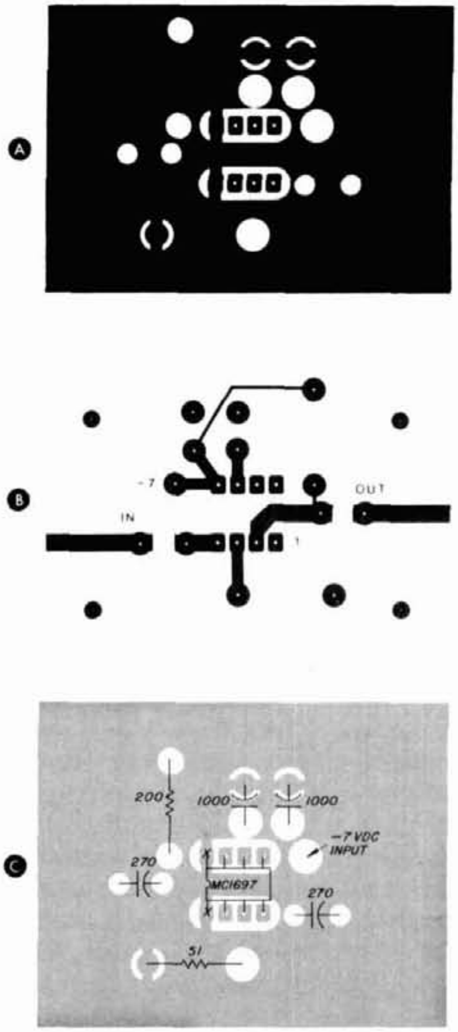

fig. 2. The top and bottom sides of the printed circuit board are shown in A and B respectively. In the parts placement diagram, C, C1, and C2 are installed on the top of the board if glass or mica capacitors are used; if chip ceramics are used, they are mounted on the foil side. The leads marked with an X are to be soldered on the ground plane side.

be brought into the box with a feedthrough bypass-type capacitor. Alternatively, a small power supply can be built in the box containing the prescaler.

operation

This prescaler module has no adjustable components and should work, Murphy willing, when it is first turned on. However, if it does not, there are several points to check. First, see that the −7 volts is present at pin 5 on the MC1697. If you have an oscilloscope, check to see that the IC is not oscillating. (A cure for oscillations is to connect a 10k-ohm resistor from pin 4 to V_{ee}.) Also check the grounds on the two bypass capacitors, pins 1 and 8 on the MC1697, and on the 51-ohm resistor. All of these ground points must be soldered to the top of the printed circuit board.

Once the prescaler is built and operating you will naturally want to check it to find the maximum operating frequency. One way of doing this involves finding a signal generator or source that covers the 1- to 2-GHz range. Attach the prescaler and counter to the generator. Start with the generator set near 1 GHz and verify that the counter reads approximately 250 MHz.

Slowly increase the frequency of the signal generator and watch the counter display. Somewhere between 1500 and 1700 MHz (about 375 to 425 MHz on the display) the indicated frequency will suddenly jump downward. This shows that the divider is no longer able to operate normally. The point at which the display jumps to a radically different value indicates the maximum input frequency at which your prescaler will operate.

There is another test that will help locate the critical upper frequency. While slowly increasing the input signal frequency, watch the least significant digits on the display of the counter. As the maximum operating frequency is approached these digits will become erratic, indicating that the phase noise on the prescaler's output is becoming significant. This flickering occurs a few megahertz below the cutoff frequency.

If you wish to vary the power supply voltage slightly, you may be able to find a point at which the maximum operating frequency is increased. However, I found that presetting the supply to −7 volts is nearly optimal, and that adjusting the voltage never raised the upper frequency limit by more than a small amount.

Another good test of the prescaler is to count a signal whose frequency is near the maximum limit for the counter itself, note the frequency displayed, and then count the same signal using the prescaler. If the counts (after multiplying by four) are very nearly the same, the prescaler is probably operating correctly.

The most serious shortcoming of this prescaler module is its limited dynamic range. Typically, the input signal must be in the range of 200 to 1600 mV p-p for proper operation. The dynamic range could be extended by the use of an input stage of amplification and limiting. The problem with this approach is that the amplifier stage must be broadband, covering at least 600 to 1500 MHz; such amplifiers are not easily built. Avantek[2] has published an application

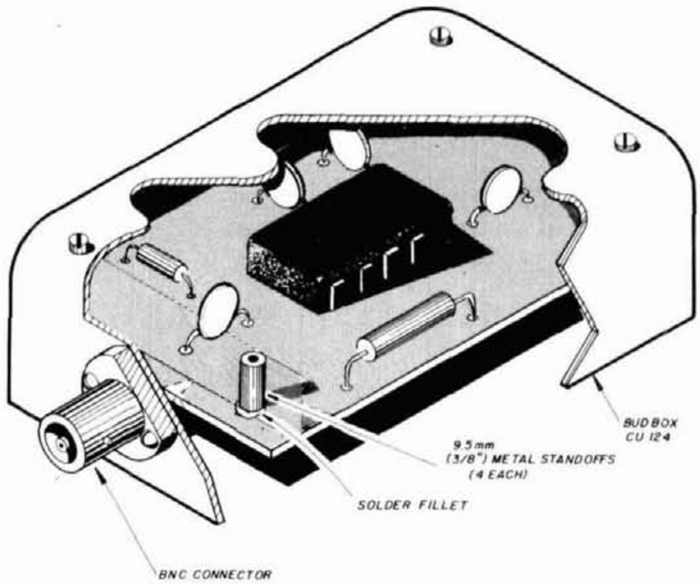

fig. 3. Mounting details for the divide-by-four prescaler. The center pin of the BNC connectors is soldered to the trace on the printed circuit board.

note which covers the construction of a 40-dB dynamic range preamplifier covering 25 to 1000 MHz. Using thin-film hybrid amplifier modules to achieve its performance, its principal shortcoming lies in the cost — more than $120.

The simple prescaler described here, even without a preamplifier, is a useful test instrument. If the input signal's amplitude is outside the operating range, the prescaler does not produce spurious outputs — it simply stops operating. As long as the input signal source can provide at least 1 mW, some attenuator level can be found which will permit operation of the scaler. Several versions of this prescaler module have been built. All functioned without problems and all had a maximum operating frequency of at least 1500 MHz.

references

1. Chris Cogburn, K5VKQ, "How to Design Regulated Power Supplies," *ham radio*, September 1977, page 58.

2. "A 1-GHz Prescaler Using GPD Series Thin-Film Amplifier Modules," *Avantek Application Note ATP 1036*, Avantek Corporation, 3175 Bowers Avenue, Santa Clara, California 95051.

ham radio

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An Exclusive State of the Art

Hy-Gain's 3750 covers all amateur bands from 1.8 through 30 MHz. It utilizes advanced Phase-Lock-Loop circuitry, dual gate Mos Fet's at all critical RF amplifier and mixer stages, a narrow band SSB crystal filter and a 50 kHz T-notch filter.

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More Details? CHECK — OFF Page 150

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The Bible tells us in Psalms 37:4, 5; "Delight thyself also in the Lord, and He shall give thee the desires of thine heart. Commit thy way unto the Lord; trust also in Him, and He shall bring it to pass". (KJV)

Jesus tells us in the Gospel of John that He is the way, the truth and the life. If we believe this, follow His teachings and obey His commands, we may ask any request of Him and it will be granted. He has told us this so we will be filled with His joy.

His way for our life will fulfill our desires and solve the complex and confusing problems of this life. Jesus said, "I am come that they might have life, and that they might have it more abundantly". John 10:10b (KJV)

God's plan for our life makes us a complete person through Jesus Christ. Please accept His love and have a blessed Christ-centered holiday season.

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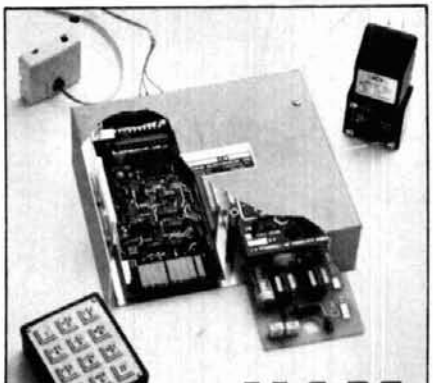

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NEW products

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vlf-hf communications receiver

The new HF1030 communications receiver introduced by Communications Product Corporation is a fully synthesized, tunable, solid-state receiver that covers the frequency range from 10 kHz to 30 MHz in 10-Hz steps. Designed by Dr. Ulrich L. Rohde, DJ2LR, well known to *ham radio* readers for his many contributions to modern high-frequency receiver design, the HF1030 has many unusual features not previously available. Included are frequency preset, remote control, and remote programmability — features which permit the receiver to be used in computer-based communications systems. Communications Product Corporation is presently working with Tektronix on a system for military applications which allows a Tektronix 4051 computer to store and decipher Morse code and RTTY transmissions. The Tektronix 4051 computer is also capable of remotely controlling and programming the HF1030 via an interphase card and IEEE bus, allowing simultaneous control of up to 30 separate HF1030 receivers.

Other features of the HF1030 include provision for all reception modes (a-m, ssb, and CW), 455-kHz output for FSK demodulation (FSK filter built in), selectable i-f bandwidths, electronic bandpass tuning, and excellent immunity to strong-signal overload. The two selectable tuning speeds of the optical shaft encoder (1800 Hz and 180 Hz per 360° dial revolution) give the feel and smoothness of an analog VFO in quasi-continuous tuning with absolutely backlash-free performance while retaining the accuracy and stability of the internal or external frequency standard. No separate MHz control is required — the frequency may be preset with the thumbwheel switches, then returned to VFO control; the fast-switching synthesizer permits frequency jumps of 30 MHz in less than 10 milliseconds.

All internal frequency control, including the BFO, is derived from a master crystal standard; worst case frequency stability for the oscillator is 1 Hz/C°. For more demanding stability requirements, there is provision for an external rubidium or cesium standard. The fully synthesized BFO provides a ±5 kHz tuning range. Almost all frequency synthesizer circuitry is based on CMOS circuits; this reduces power consumption to less than 2 watts and enhances reliability because of heat reduction. Spectral purity of the synthesizer is improved because of the great reduction of digital noise. The CMOS memories and standby battery provide storage of frequency data for at least one year in the event of power failures; upon restoration of power, the receiver will return to the previously tuned frequency.

The HF1030 receiver uses a specially developed, high-level double-balanced mixer with monolithic hot-carrier diodes, which is terminated in a low-noise amplifier with heavy feedback for excellent sensitivity without any requirement for an rf amplifier. Sensitivity on CW, FSK, and ssb is typically 0.5 μV for 10 dB signal-to-noise ratio; a-m sensitivity is 2.6 μV typical for 10 dB S/N. The noise figure is less than 10 dB from 1-30 MHz. The third-order input intercept point is +20 dBm (signals separated 30 kHz); second-order IMD is −80 dB.

The first i-f of the HF1030 is at 40.455 MHz; the second i-f is at 455 kHz. To improve input selectivity the 8-pole 40.455-MHz crystal filter has a bandwidth of ±4 kHz with a 1:2 shape factor; this filter was designed specifically for low intermodulation distortion products. The five built-in 455-kHz i-f filters have been specially selected for optimum performance on ssb (2.7 kHz), CW (375 Hz), FSK (1.9 kHz), and a-m (5.8 kHz). The built-in bandpass tuning feature (controlled by the main tuning knob) is accomplished by electronically shifting the frequencies of the LO and BFO.

The very effective AGC system used in the HF1030 receiver has a minimum range of 120 dB with a threshold at 0.2 μV; there is less than 6 dB audio change with rf inputs from 1 μV to 100 mV. On ssb, the AGC attack time is 100 ms; hold time is 1.6 seconds and release time is 50 ms (typical). For a-m reception, AGC attack time is 10 ms; release time is 35 ms.

Other features of the HF1030 include a very effective squelch system which discriminates between man-made noise or electrical noise and voice or other wanted transmissions. Editor-in-chief W1HR had an opportunity to compare the performance of the HF1030 against several other high-frequency receivers and reported that "It's probably the best communications receiver I've ever used; weak signals that were completely covered by noise and internal IMD in other receivers were solid copy on the HF1030. It's amazing how much of the splatter hams complain about is actually being generated right in their own receivers!"

The receiver uses modular construction with easy access to all assemblies; its compact size allows two

HF1030 receivers to be mounted side by side in a single rack panel. The set measures 8.24 inches wide, 5.22 inches high, and 14.48 inches deep (21 × 13.3 × 36.8 cm); weight is 18 pounds (8.2 kg). Total power consumption (all LED displays on) is less than 15 watts. Power requirement is 110/220 Vac, 47-400 Hz at 25 watts, or 15 to 25 Vdc at 1 amp. Further information on the HF1030 Communications Receiver can be obtained from Rohde & Schwarz Sales Company, (U.S.A.) Inc., 14 Gloria Lane, Fairfield, New Jersey 07006.

bencher keyer paddle

Bencher, Inc., of Chicago, is producing a dual-lever, iambic keyer paddle that has a lot of features you'll find helpful for ease and reliability in CW operation. Some of these features are dual adjustment of spring tension on the paddles to match the "feel" to your fist; precision contact adjustments with a set-screw locking feature so the adjustment stays where you want it; and a heavy steel base, equipped with nonskid feet so you don't have to chase the paddle around the desk top.

The contact points are solid silver, for a lifetime of flawless keying. Nylon bearings are used on the keying shafts, and the bearings float on machined brass fittings for a self-aligning action. Spring tension prevents play and sloppiness, eliminating contact bounce and backlash.

The Bencher Ultimate Paddle looks good and works well. The frame, posts, and bearing ring are all precision machined from solid brass stock, polished and plated for durability and first-rate appearance.

You can obtain the model BY-1 (standard black-finish base) for

$39.95; or the BY-2 (polished chrome-plated base) for $49.95. Bencher will quote the price of a Gold Plated Presentation model (BY-3) upon request. The paddles are available at selected dealers, or direct from Bencher, Inc., 333 West Lake Street, Chicago, Illinois 60606.

Palomar synthesized transceiver

The Palomar PTR-130K is the first completely multi-functional transceiver ever made available to the public. It incorporates some of the most advanced features of logic technology. The PTR-130K is a miniaturized mobile transceiver capable of operating with 100-Hz resolution from 100 kHz to 30 MHz, in all modes of transmission and reception (ssb, CW, fm, and a-m). Instant frequency selection is available at the touch of a finger.

The transceiver has a built-in rf compressor to increase effective transmitter output by 12 dB (16 times). Two cascaded Collins mechanical filters in the receiver provide superior selectivity, with a typical shape factor of 1:1.25 (3 dB/60 dB). These, and many other features, are packed into a small, 16.5 × 6.3 × 20 cm (6-1/2 × 2-1/2 × 8-inch) package. Power output is 5 watts on a-m, 12 watts on other modes.

The PTK-130K is the latest in a long line of innovative electronics products from Palomar Electronics Corporation, 665 Opper Street, Escondido, California 92025.

"Band-Aids" Handbook

James Dersch, WDØAJE, has just published *Band-Aids*, which he calls the "Radio Amateur Operator's Handiest Handbook." If you've ever

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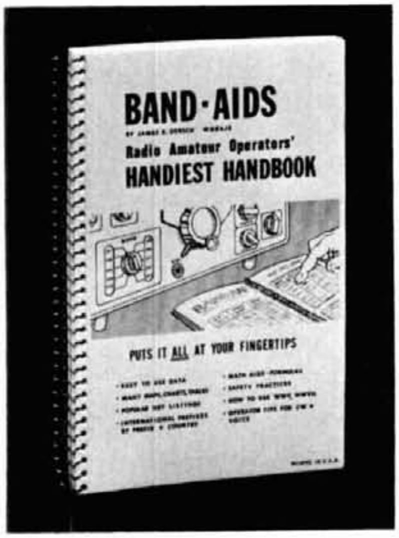
had to frantically search for some bit of information in the midst of a QSO or contest, you will agree with that description.

Band-Aids is a spiral-bound booklet full of handy information in a multitude of categories. It contains many operating aids that the amateur will find useful — perhaps essential — during the many activities he pursues. It has maps and lists for such things as the Worked All States award (WAS), time-zone information, abbreviations, IACO word list (phonetic alphabet), OSCAR frequencies, U.S. 160-meter allocations, W1AW broadcast schedule, third-party agreement information, amateur network listings and Q-signals, international prefix lists, and many other bits and pieces that are always needed at your fingertips but manage to get lost at the wrong time.



In the nonoperating category, *Band-Aids* contains useful material on metric conversions, resistor (and other component) color coding, schematic-diagram symbols, telephone touch-tone frequencies, vswr nomograph, some useful formulas, and much more.

Band-Aids contains 110 pages, and measures 14 × 21 cm (5-1/2 × 8-1/2 inches). The spiral binding allows it to lie flat, and the compact size allows it to fit comfortably on even the smallest of operating positions. You can obtain one from Ham

Radio's Communications Bookstore, Greenville, New Hampshire 03048, for $7.95, plus 40 cents shipping and handling. Order CC-BA.

tunable sub-audible tone encoder

Vega introduces its new subminiature, subaudible tone encoder for hand-held as well as routine mobile radio applications. The Model 188 is ideal for one-way, tone-protected applications where all mobiles are equipped to transmit, but not to receive, tone. The tone frequency is continuously field tunable and no modifications or elements are required to establish the tone frequency. Extremely stable, the Model 188 maintains frequency and level accuracy over the temperature and voltage variations found in mobile radio environments. The unit measures 2 × 2 × 1.3 cm (0.8 × 0.85 × 0.52 inch) and is backed by a three-year warranty. It is available for quick delivery.

For further information, contact VEGA, 9900 Baldwin Place, El Monte, California 91731.

amateur antenna catalog

A new catalog covering their comprehensive line of mobile and base station antennas for amateur radio applications has just been issued by Antenna Incorporated, Cleveland, Ohio. The 8-page catalog provides detailed descriptions and complete electrical and mechanical specifications on some 4 dozen ham antennas, including 10 meter, 6 meter, 2 meter, 3/4 meter, and 1-1/4 meter types.

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Thirteen types of mountings are available, according to Randall J. Friedberg, vice president and sales manager. The antennas are designed for 100, 150, and 200-watt power ranges. The mobile units are designed for temporary or permanent installation on all types of vehicles, Friedberg said.

All of the antennas and accessories described in the 8-page booklet are manufactured in the United States and of highest quality materials to assure dependable performance.

Copies of the catalog are available free on request from Antenna Incorporated, 26301 Richmond Road, Cleveland, Ohio 44146.

CW/ssb active filter

MFJ Enterprises has introduced two new CW and ssb active filters. The top-of-the-line model is called the MFJ-721 Super Selector CW/ssb Filter. It has a 2-watt audio amplifier, switchable noise limiting, and an input selector switch for two rigs.

The CW filter is an 8-pole active filter (4 cascaded stages) centered nominally at 750 Hz. It has four selectable bandwidths, 180, 150, 110, and 80 Hz. In the 80-Hz position, the response is at least 60 dB down one octave from the center frequency. It drastically reduces noise and provides up to 15 dB improvement in signal-to-noise ratio.

With a pair of stereo headphones, simulated stereo reception directs the narrow, filtered signal to one ear and the unfiltered signal to the other. The ears and brain then reject the interference, but allow off-frequency calls to be heard.

The ssb filter dramatically improves

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readability by optimizing the audio bandwidth to reduce sideband splatter, remove low- and high-pitched QRM, hiss, static crashes and background noise, and eliminate 60- and 120-Hz hum.

A self-adjusting automatic peak clipper is provided for ssb. For CW, another clipper removes background noise that is smaller in amplitude than the signal. The MFJ-720 plugs into the phone jack and drives a speaker or phones with 2 watts of audio. The size is 12.7 × 5 × 15.2 cm (5 × 2 × 6 inches). It requires 9 to 18 Vdc; an optional ac adapter is available. The price is $59.95.

The MFJ-720 Deluxe Super CW Filter uses the same 8-pole active filter as the MFJ-721. The frequency determining components are hand selected for a response within one Hz of the nominal 750-Hz center frequency. The low-Q cascaded design minimizes ringing. A self-adjusting peak noise limiter is also built in.

The MFJ-720 plugs into the phone jack and drives the speaker or phones. The size is 10.2 × 5 × 15.2 cm (4 × 2 × 6 inches); it requires 9 to 18 Vdc. An optional ac adapter is available. The price is $44.95.

The MFJ-721 Super Selector CW/ssb Filter and the MFJ-720 Deluxe Super CW Filter are available from MFJ Enterprises; they both have a 30-day, money-back trial period. If you are not satisified, you may return the filters within 30 days for a full refund (less shipping). MFJ also provides a one-year unconditional warranty.

To order, call toll free 800-647-8660, or mail your order to MFJ Enterprises, Post Office Box 494, Mississippi State, Mississippi 39762. Include $2.00 for shipping and handling.

Chemtronics desoldering tool

Chemtronics Inc., of Hauppauge, New York, recently announced the latest addition to its popular line of solder and industrial-chemical prod-

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ucts: the D5 Desoldering Tool. This unique new product, which features Chemtronics' highly effective desoldering wick in a specially engineered, refillable dispenser tool, helps technicians remove solder more efficiently while economizing on wick use. D5 may be used alone, or as an integral part of Chemtronics' new SD5 Solder/Desolder System.

The D5 Desoldering Tool consists of a 25-mm (1-inch) clear-plastic cylinder which contains a visible supply of 152 cm (5 feet) of the company's specially formulated desoldering wick. Braid is fed to the wick through a Teflon* probe that extends from one end of the wick supply. The heat-resistant Teflon probe allows users to desolder with pinpoint accuracy and without burnt fingers, even in high-density circuitry. In addition, the D5 Tool's exclusive probe permits the user to shape or "web" the wick, providing maximum absorbency and further economizing on wick use. When the wick supply is exhausted, the user simply snaps the tool's probe into the D5 Desoldering Tool Wick Refill.

Chemtronics' D5 Desoldering Tool uses the highest-quality braid in natural copper color, which permits the user to see the absorption of solder. The braid, which meets all MIL-specs and NASA publications requirements, is treated with a pure, water-white rosin flux which is nonactivated and free from halogens and corrosive chlorides. This assures complete solder absorption without leaving harmful residue.

The pocket-sized D5 Desoldering Tool is available alone or as part of the SD5 Solder/Desolder System, where it telescopes or snaps in and out of a pound or half-pound spool of Chemtronics' solder. D5 wick refills are also available, in two diameters, allowing the D5 Tool to be economically reused for years. More information is available at Chemtronics distributors or directly from Chemtronics Inc., Solder Products Division, 45 Hoffman Avenue, Hauppauge, New York 11787.

*Teflon is a registered DuPont trademark.

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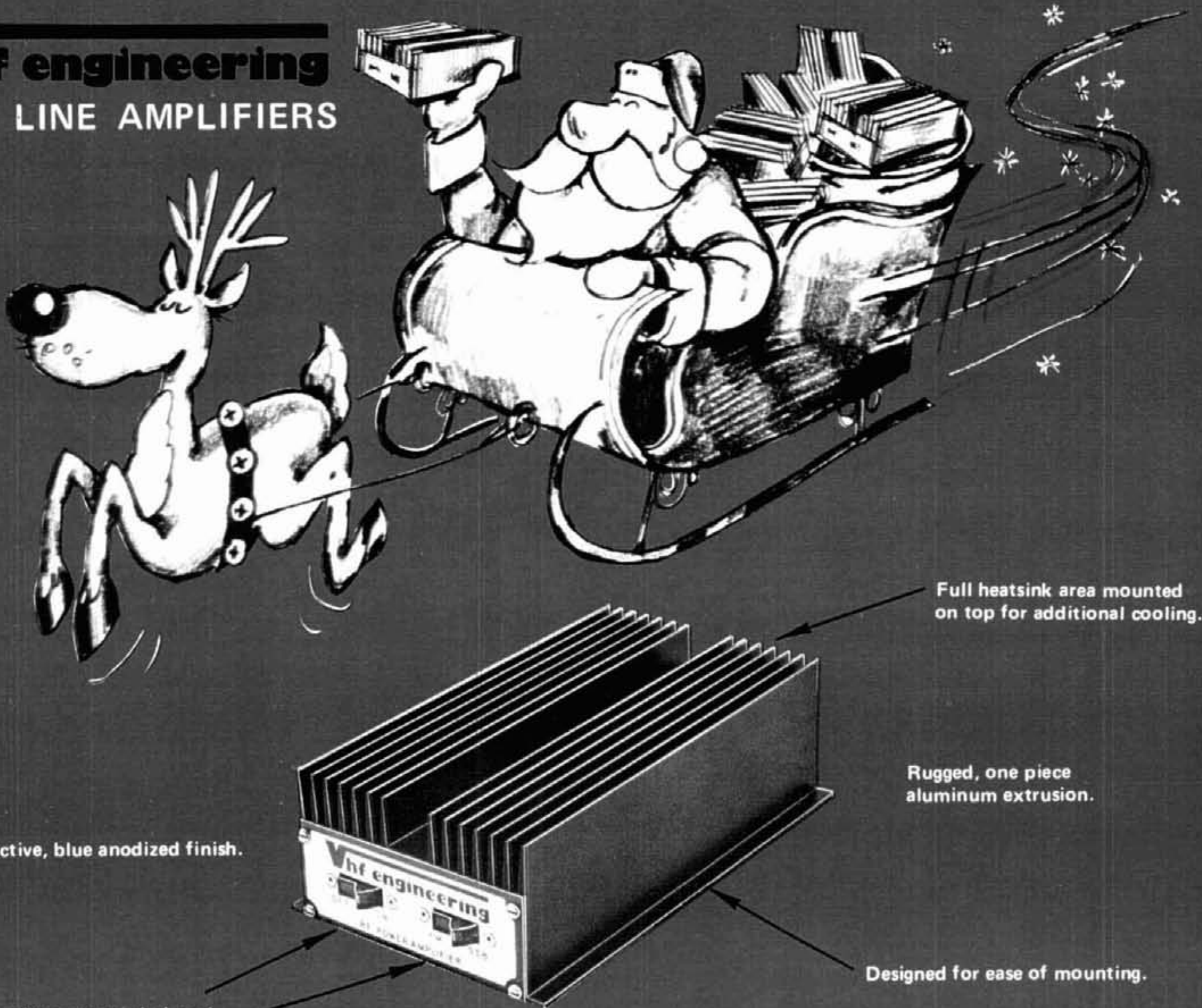

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BLC 2/70	144 MHz	CW-FM-SSB/AM	2W	70W	169.95
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BLC 30/150	144 MHz	CW-FM-SSB/AM	30W	150W	239.95
BLD 2/60	220 MHz	CW-FM-SSB/AM	2W	60W	164.95
BLD 10/60	220 MHz	CW-FM-SSB/AM	10W	60W	159.95
BLD 10/120	220 MHz	CW-FM-SSB/AM	10W	120W	259.95
BLE 10/40	420 MHz	CW-FM-SSB/AM	10W	40W	159.95
BLE 2/40	420 MHz	CW-FM-SSB/AM	2W	40W	179.95
BLE 30/80	420 MHz	CW-FM-SSB/AM	30W	80W	259.95
BLE 10/80	420 MHz	CW-FM-SSB/AM	10W	80W	289.95

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ALDA 103 is completely manufactured in the U.S.A.

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HAPPY BIRTHDAY! Now ten years fighting TVI. The RSO low pass filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada M1S 3B4.

WANTED: Regency HR-2, HR-440, AR-2 amp, SBE-450, 2 meter cavities, cash or trade Heath SB-620 Panadapter. K6KZT, 2255 Alexander, Los Osos, CA 93402.

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RECONDITIONED TEST EQUIPMENT for sale. Catalog $.50. Walter, 2697 Nickel, San Pablo, Ca. 94806.

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HOMEBREWERS: Stamp brings component list. CPO Surplus, Box 189, Braintree, Mass. 02184.

ELECTRONIC BARGAINS, CLOSEOUTS, SURPLUS! Parts, equipment, stereo, industrial, educational. Amazing values! Fascinating items unavailable in stores or catalogs anywhere. Unusual FREE catalog. ETCO-012, Box 762, Plattsburgh, N.Y. 12901. SURPLUS WANTED.

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ELECTRONIC EQUIPMENT HOTLINE is a classified advertising newsletter for professional, industrial, and surplus electronic equipment. Subscriptions $8/year, ads 50¢/word. P.O. Box 4768, Dept HH, Panorama City, CA 91412.

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Coming Events

SOUTH BEND, INDIANA Hamfest Swap & Shop, January 7, 1979, first Sunday after New Years Day at New Century Center downtown by river on U.S. 31 Oneway North across from St. Joseph Bank Building. Half acre in one large room at ground level. Tables $3 each. Food service, automobile museum and Art Center in same building. Four lane highways to door from all directions. Talk-in Freq: 146.52-52, 13-73, 34-94; 147.99-39, 93-33, 84-24, 69-09.

ROCHESTER Hamfest & NY State ARRL Convention, May 25-27. Add your name to mailing list. Send QSL to Rochester Hamfest, Box 1388, Rochester, NY 14603. Phone (716) 424-1100.

WORKSHOPS: An expanded and new series of four lecture/laboratory workshops on microcomputer data acquisition, instrumentation and measurement systems are being given by the authors of the popular Bugbooks. Course dates are November 27 to December 2, 1978 and December 18-21, 1978. For more information, contact Dr. Linda Leffel, Center for Continuing Education, Virginia Polytechnic Institute and State University, Blacksburg, VA 24061. (303) 961-5241.

TEN TEN INTERNATIONAL Net Winter CW QSO Party. December 16-17. 0000Z-2400Z the next day. (48 hour period). All contacts below 28.5 MHz. Exchange: Name, QTH, 10-x number (if you have one). Scoring: 1(one) point per contact, 1(one) point if with a 10-x member, 1(one) point if with a novice or technician. Novices sign/N, Technicians sign/T. (maximum score per completed contact is 3 points). Be sure to specify date and time of each contact on your logs. Give chapter for chapter credit. Send your name, QTH, call, and 10-x number along with all logs to: Daniel Rubin WA1ZXB, 12 Princeton Street, Danvers, MA 01923, no later than January 20, 1979. Awards given.

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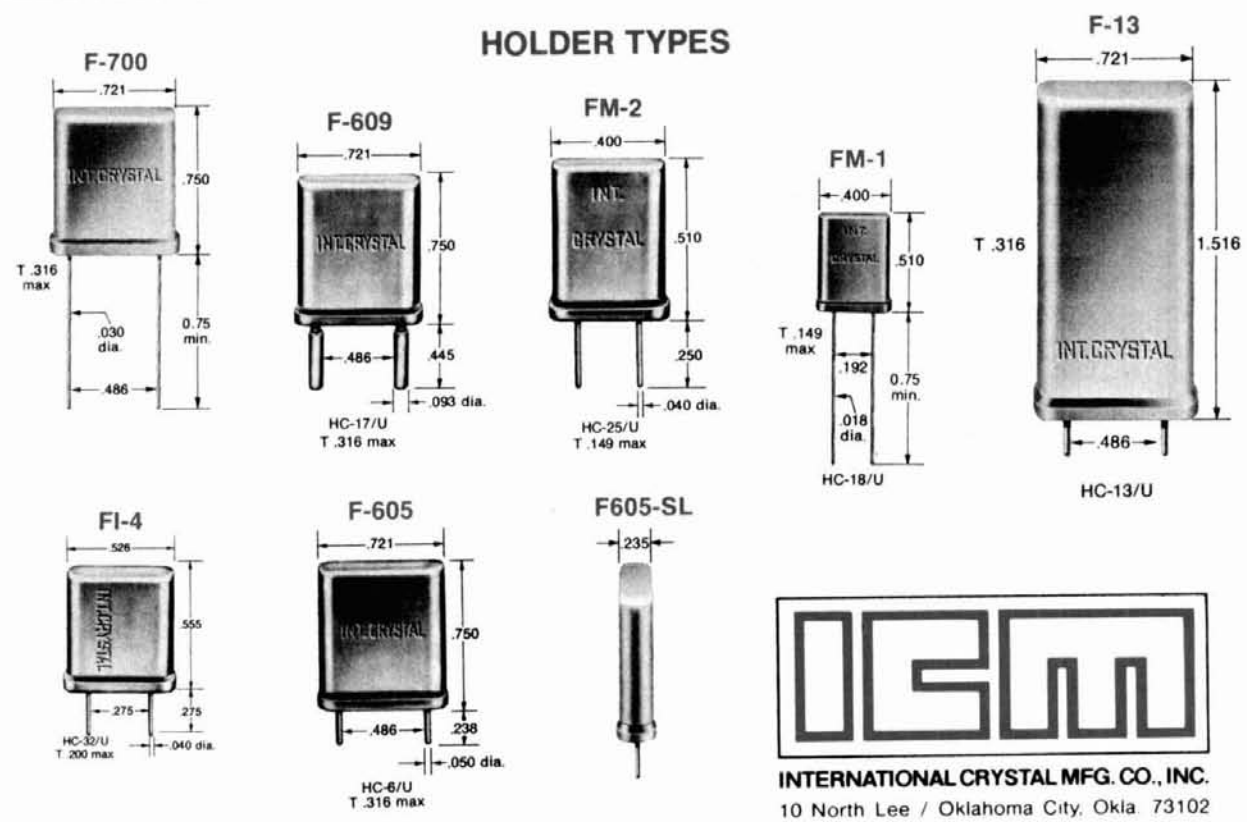

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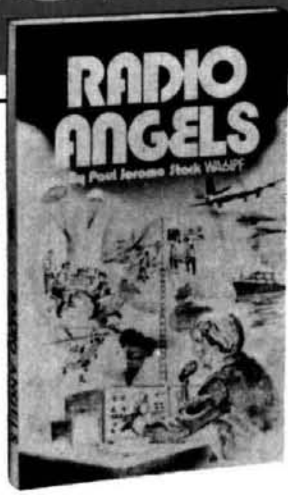

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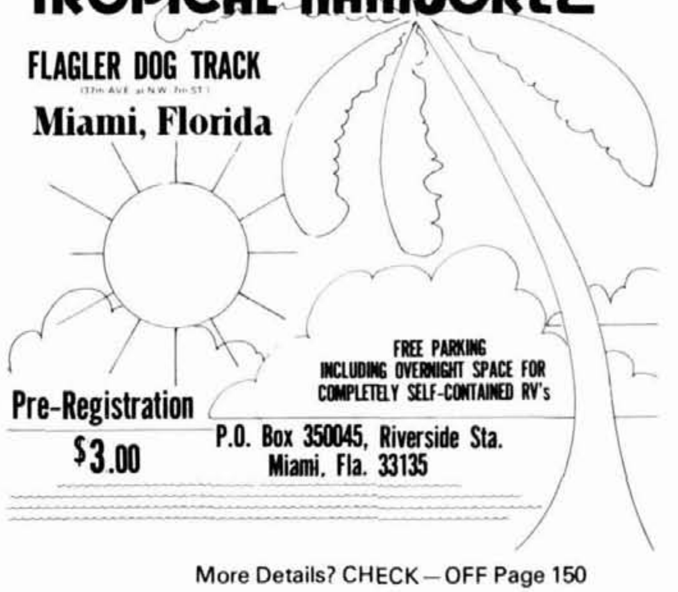

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1N914	100v	10mA	.05
1N4005	600v	1A	.08
1N4007	1000v	1A	.15
1N4148	75v	10mA	.05
1N4733	5.1v	1 W Zener	.25
1N753A	6.2v	500 mW Zener	.25
1N758A	10v	"	.25
1N759A	12v	"	.25
1N5243	13v	"	.25
1N5244B	14v	"	.25
1N5245B	15v	"	.25

SOCKETS/BRIDGES

8-pin	pcb .20	ww	.35
14-pin	pcb .20	ww	.40
16-pin	pcb .20	ww	.40
18-pin	pcb .25	ww	.75
22-pin	pcb .35	ww	.95
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2N3055	NPN 15A 60v	.50
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4002	.20
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4009	.35
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4011	.20
4012	.20
4013	.40
4014	.75
4015	.75
4016	.35
4017	.75
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4020	.85
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4043	.50
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4046	1.25
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4050	.45
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4071	.25
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4082	.30
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MC 14419	4.85
4511	.95
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7400	.10	7473	.25	74176	.85	74H72	.35	74S133	.40
7401	.15	7474	.30	74180	.55	74H101	.75	74S140	.55
7402	.15	7475	.35	74181	2.25	74H103	.55	74S151	.30
7403	.15	7476	.40	74182	.75	74H106	.95	74S153	.35
7404	.10	7480	.55	74190	1.25			74S157	.75
7405	.25	7481	.75	74191	.95	74L00	.25	74S158	.30
7406	.25	7483	.75	74192	.75	74L02	.20	74S194	1.05
7407	.55	7485	.55	74193	.85	74L03	.25	74S257 (8123)	1.05
7408	.15	7486	.25	74194	.95	74L04	.30		
7409	.15	7489	1.05	74195	.95	74L10	.20	74LS00	.20
7410	.15	7490	.45	74196	.95	74L20	.35	74LS01	.20
7411	.25	7491	.70	74197	.95	74L30	.45	74LS02	.20
7412	.25	7492	.45	74198	1.45	74L47	1.95	74LS04	.20
7413	.25	7493	.35	74221	1.00	74L51	.45	74LS05	.25
7414	.75	7494	.75	74367	.75	74L55	.65	74LS08	.25
7416	.25	7495	.60			74L72	.45	74LS09	.25
7417	.40	7496	.80	75108A	.35	74L73	.40	74LS10	.25
7420	.15	74100	1.15	75491	.50	74L74	.45	74LS11	.25
7426	.25	74107	.25	75492	.50	74L75	.55	74LS20	.20
7427	.25	74121	.35			74L93	.55	74LS21	.25
7430	.15	74122	.55			74L123	.85	74LS22	.25
7432	.20	74123	.35	74H00	.15			74LS32	.25
7437	.20	74125	.45	74H01	.20	74S00	.35	74LS37	.25
7438	.20	74126	.35	74H04	.20	74S02	.35	74LS38	.35
7440	.20	74132	.75	74H05	.20	74S03	.25	74LS40	.30
7441	1.15	74141	.90	74H08	.35	74S04	.25	74LS42	.65
7442	.45	74150	.85	74H10	.35	74S05	.35	74LS51	.35
7443	.45	74151	.65	74H11	.25	74S08	.35	74LS74	.35
7444	.45	74153	.75	74H15	.45	74S10	.35	74LS86	.35
7445	.65	74154	.95	74H20	.25	74S11	.35	74LS90	.55
7446	.70	74156	.70	74H21	.25	74S20	.25	74LS93	.55
7447	.70	74157	.65	74H22	.40	74S40	.20	74LS107	.40
7448	.50	74161	.55	74H30	.20	74S50	.20	74LS123	1.00
7450	.25	74163	.85	74H40	.25	74S51	.25	74LS151	.75
7451	.25	74164	.60	74H50	.25	74S64	.15	74LS153	.75
7453	.20	74165	1.10	74H51	.25	74S74	.35	74LS157	.75
7454	.25	74166	1.25	74H52	.15	74S112	.60	74LS164	1.00
7460	.40	74175	.80	74H53J	.25	74S114	.65	74LS193	.95
7470	.45			74H55	.20			74LS367	.75
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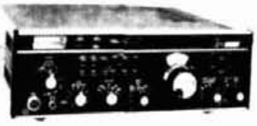
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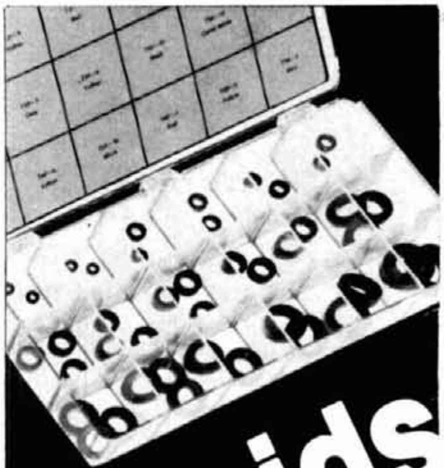

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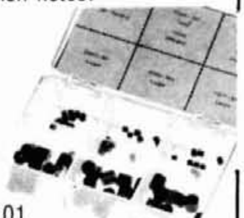

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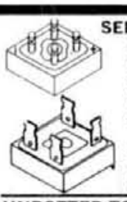

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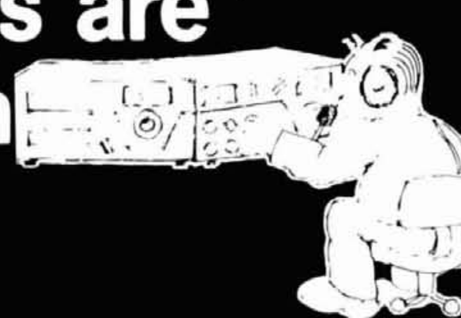

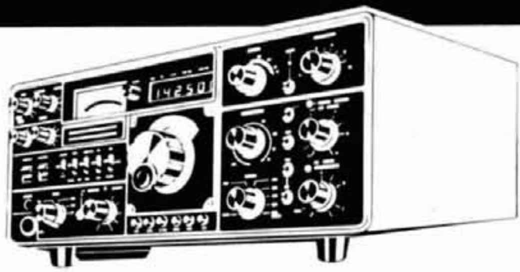

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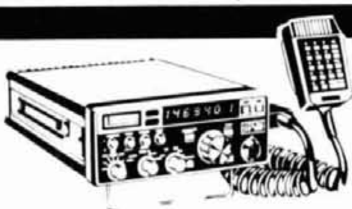

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Keyboard mic for remote input, scanning control. 800 channels, freq. 144-148 MHz.

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SP-101PB speaker/phone patch console

Response speaker 300-3000 Hz, individual gain controls, full VOX phonepatch op., input 4 or 600 ohms, output 600 ohms or high impedance.

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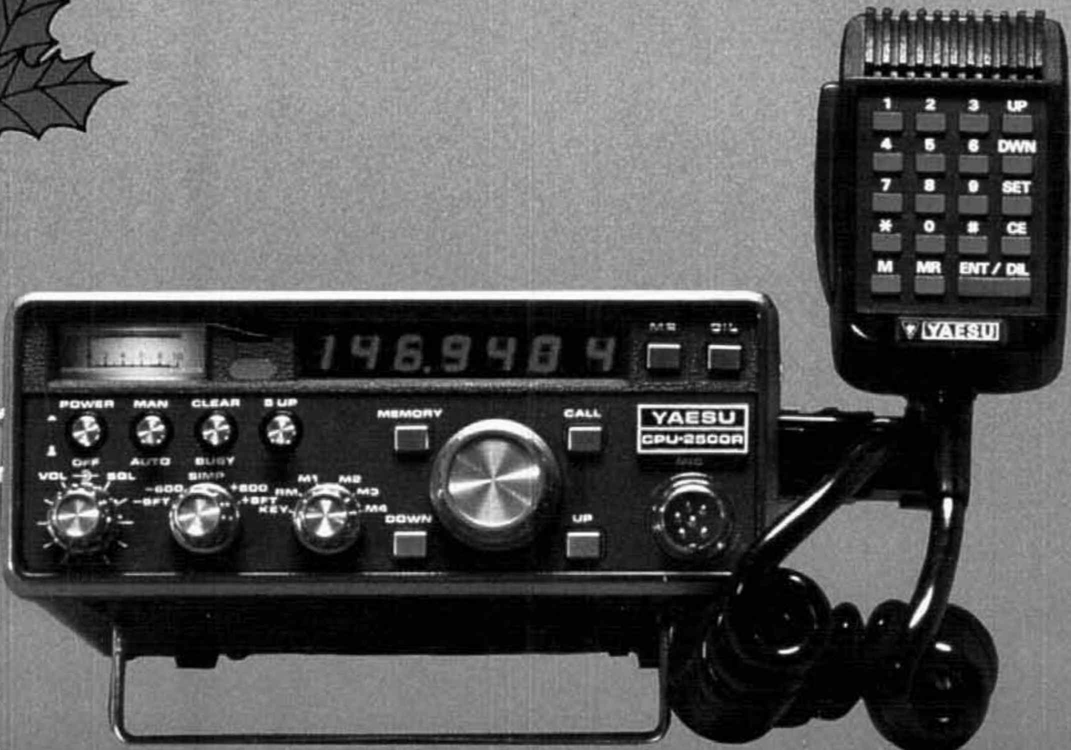
More Details? CHECK—OFF Page 150

THE CHRISTMAS GIFT OF A LIFETIME!

AESU ANNOUNCES THEIR SENSATIONAL COMPUTER AGE CPU-2500R/K 2-METER 25 WATT TRANSCEIVER



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The CPU-2500R/K is a space age radio for discriminating amateurs utilizing the latest solid state techniques and it's on your dealer's shelf today!

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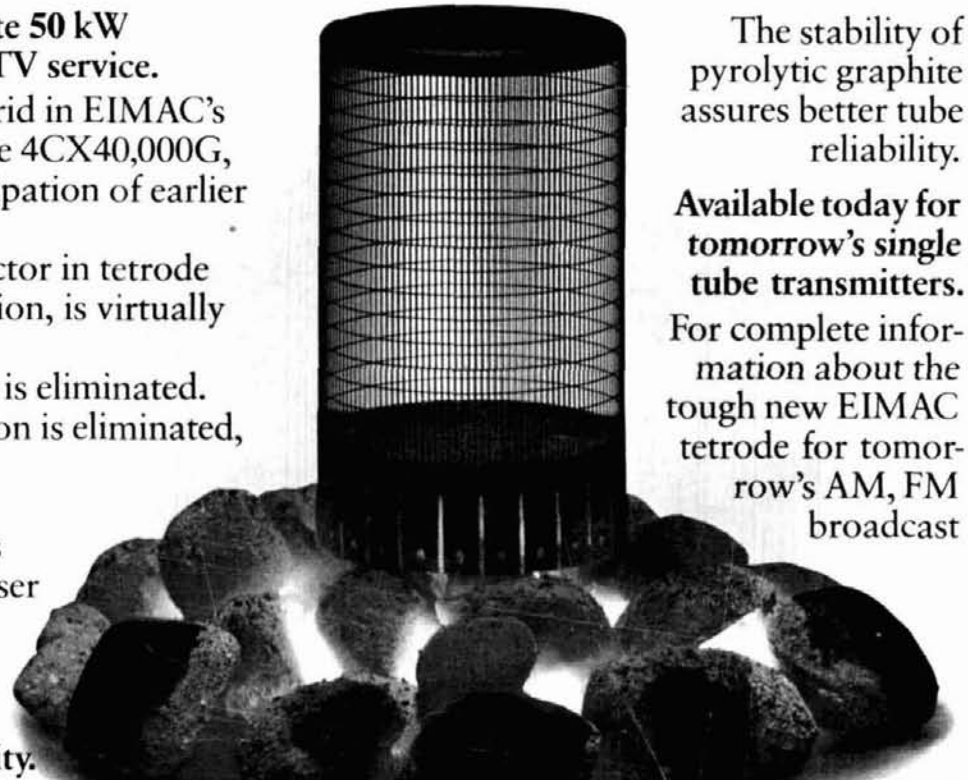
EIMAC's new Pyrogrid can run hotter so your transmitter can run better.



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